Analysis and Design of Single-Sided, Slotted AMM Axial-Field Permanent Magnet Machines

by

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Chapter 1

Introduction

1.1 Background

The introduction of variable speed motor drives in industrial applications offers higher efficiency over a wide range of operating conditions. Since 1990 the number of electrical motors used in automobiles have increased significantly as reported in [1]. Furthermore, there is a trend in automotive applications to increase the operating speed of electric motors in order to reduce their size, which leads to weight and cost reductions. Although using variable-speed drives can save significant energy, the losses in electrical machines constrain the overall efficiency of the drive, specifically at high speed operation. Moreover, the output characteristics of electrical machines and their production cost also have a significant effect on their selection.

The two major sources of losses in electrical machines are copper loss in the stator winding and iron loss in the magnetic core. A better utilisation of copper in the magnetic circuit design can reduce copper losses such as using a concentrated-winding design with short end windings. Iron losses in machines however, are strongly affected by the frequency of excitation, which is proportional to the speed of the machine and so are particularly important at high speeds. In order to reduce iron losses an attractive approach is to utilise new magnetic materials as the core of the machines, which can offer lower losses than the conventional materials.

Despite the rapid advancements in power electronics, the basic construction of electric motors has remained unchanged for over a hundred years, which primarily utilise silicon iron (SI) as the core material. Although SI based electrical machine manufacturing methods are well established and simple, SI has significant iron losses specifically in high frequency variable-speed motor drive applications. However, recent developments offer two emerging materials for magnetic cores, amorphous magnetic materials (AMM) and soft magnetic composite (SMC). These materials each offer unique characteristics that can be utilised to improve the performance of rotating electric machines.
Chapter 1 Introduction

Permanent magnet (PM) machines are well known for their high power density and efficiency. Recent developments of advanced power electronics and affordable permanent magnets have attracted interest in various applications based on PM motor topologies. Axial field PM (AFPM) machines in particular are gaining popularity due to their short axial length and high torque, power density and efficiency [2–4]. In addition, its potential for low cogging torque and operating noise make it a good candidate in some applications such as ship propulsion drives [5] and direct drives in elevators [6].

It was investigated that AFPM machines are well suited to AMM due to their use of a tape-wound structure. In addition, the industry partner of this project, Glassy Metal Technologies (GMT), has developed a commercial viable and accurate cutting technique that is suitable for tape-wound structure. Therefore, this work aims to utilise the above mentioned advantages of AMM material and the benefits of an AFPM structure in rotating PM electrical machines. The principal aim of this research work is to design and construct cost effective and highly efficient PM machines based on AMM to be used in generally high speed motor or generator applications.

1.2 Motivation and Aim

The motivation for this research work is summarised as follows:

- Increased awareness of the importance of energy conservation and environmental issues.

- Requirements for high efficiency, low cost and low emissions in automobiles have driven the development of electrical machines using innovative materials and improved design techniques.

- A means to increase the efficiency of electrical machines by utilising alternative magnetic materials that offer lower iron losses than conventional silicon iron materials.

- Utilising an improved AMM material cutting method using abrasive water jet cutting techniques which would significantly reduce the manufacturing and handling cost of utilising AMM in axial-field rotating electrical machine applications.
Chapter 1

Introduction

The principal aim of this project is to utilise the AMM cutting and handling method developed by the industrial partner to develop cost effective and highly efficient AMM permanent magnet machines to be used in low cost, high speed automotive applications. Some of the targeted applications are high-speed direct-drive motors in electric vehicle, generator, vacuum cleaner and router.

Therefore, the project investigates axial-field machines utilising slotted AMM stators and concentrated windings. This involves motor design and optimisation, finite-element simulation, construction and testing of prototype machines. In addition, the project provides a direct comparison of the iron loss and performance of machines using AMM and SMC with a conventional silicon iron machine.

1.3 Outline of Thesis

In this thesis, the background of AMM and AFPM machines including material properties, topologies and fabrication methods are first described in Chapter 2. This includes discussion on the benefits and current limitations of applying AMM in rotating machines which leads to the proposal of cut AMM AFPM machines based on the cutting technique developed to be investigated in this work.

In order to quantify and compare the magnitude of the iron losses of the different materials discussed in Chapter 2, an extensive series of experimental iron loss measurements and finite element (FE) simulation were conducted in Chapter 3. The analysis and experimental results on the slotted and non-slotted cores of various materials are presented.

In the initial phase of the research, a small size (32mm diameter) three-tooth AMM stator designed and cut by the industrial partner for a targeted application was provided. In Chapter 4, a simple analytical approach was utilised to design the winding for a three slot and two pole configuration. 3D finite-element analysis (FEA) was also utilised to model, analyse and verify the design. In addition, extensive experimental tests were conducted to obtain the influence of the rotor magnet design and airgap length, and to provide the performance characteristics of the AFPM motor, including a direct comparison with SMC material. Even though the AMM motor is not an optimised design, this chapter focused on the concept demonstration and the feasibility of AMM AFPM machines which leads to a more detailed analysis and design in the following chapters.
In Chapter 5, the analytical approach is further extended to determine the stator size, slot size, magnet thickness and slot and pole combination selection for a larger size (110mm outer diameter non-slotted core sample provided by the industrial partner) AMM AFPM machine with concentrated stator windings and a surface-mounted magnet rotor. The size of the stator was chosen based on width of the AMM ribbon, AMM cutting limitation, the available rotor and the laboratory testing setup. A detailed discussion on the design variables including the optimisation of the design parameters are also presented in the chapter. This includes analytical calculation of the machine’s parameters, output torque, open-circuit losses and the design process for the prototype is demonstrated based on the proposed guidelines. 3D FEA is utilised to verify the analytical models.

In Chapter 6, a detailed discussion on the 3D FE modelling, optimisation of meshing and computation times are presented. In addition, the simulated results are compared with experimental results to examine the accuracy of the 3D modelling of the AMM AFPM machine utilised in Chapter 5.

The designed 110mm AMM AFPM machine in Chapter 5 was constructed and a comprehensive experimental results were performed. In Chapter 7, the experimental results are presented and the analytically calculated and FE simulated results are also included where available. This includes back-EMF profiles, cogging torque, axial force and loss characteristics. Furthermore, a direct comparison was performed using identical size SI and SMC based stators. The chapter also includes investigation of the machine’s parameters and performance characteristics with different airgap lengths, pole number and magnet materials.

The identified high open-circuit losses in Chapters 4 and 7 were further investigated in Chapter 8 based on the larger size machine. Additional slotted and dummy cores including a double-sided rotor were constructed for the analysis. In this chapter, the loss components separation was performed based on 3D FEA and experimental results.

Lastly, Chapter 9 summarises the major contributions of this research work and proposes further studies on AMM-based AFPM machines.
Chapter 2

Axial-Field Permanent Magnet Machines

This chapter introduces the utilisation of Amorphous Magnetic Materials (AMM) to improve the performance of rotating machines which is the main aim of this work. In order to justify the choice of AMM, a comparison with conventional silicon iron (SI) and soft magnetic composite (SMC) materials is conducted. This includes the properties, benefits and limitations of these materials in electrical machine design. In this chapter, an overview of axial-field permanent magnet machines including their features, topologies, materials and fabrication.

AMM offers extremely low iron losses, specifically at higher supply frequencies, which makes it a good candidate for high-efficiency and variable-speed motor applications. However, due to the handling and cutting limitations, AMM has not been widely utilised in rotating electrical machines. Although a few studies have reported utilising AMM in motor applications, these studies were based on costly cutting techniques or uncut versions of AMM. A commercially viable cutting technique was recently developed by the industrial partner. It is thus now practical to cut the AMM ribbons to produce stators in particularly for axial-field stator designs.
2.1 Introduction

Despite rapid advancements in the area of power electronics, the basic construction of electric motors has remained unchanged for over a century years. Most electrical machines utilise silicon-iron (SI) laminations as the stator core material. However, the laminations are limited to two-dimensional (2D) flux paths and SI has high iron loss specifically at high frequency. Recently, two materials are emerging as alternative magnetic materials: amorphous magnetic materials (AMM) which offer lower iron losses, and soft magnetic composite (SMC) allowing three-dimensional (3D) flux paths and increasing design flexibility. Table 2.1 provides a summary of the principal magnetic properties of three commercially available magnetic materials: SI, SMC and AMM.

<table>
<thead>
<tr>
<th></th>
<th>SI Transil 350-50-A5</th>
<th>SMC Somaloy 500+0.5% Kenolube 800MPa, 500°C</th>
<th>AMM 2605SA1 annealed</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnetic flux density, $T@5000A/m$</td>
<td>1.66 $(50Hz)$</td>
<td>1.45 $(500Hz)$</td>
<td>1.57 $(60Hz)$</td>
</tr>
<tr>
<td>Stacking factor, %</td>
<td>90-95 solid</td>
<td>60-80</td>
<td>60-80</td>
</tr>
<tr>
<td>Unsaturated relative permeability</td>
<td>5,000</td>
<td>500</td>
<td>600,000</td>
</tr>
<tr>
<td>Electric resistivity, $\mu\Omega cm$</td>
<td>49</td>
<td>10,000</td>
<td>130</td>
</tr>
<tr>
<td>Specific core losses, at $50Hz$ and $1T$, W/kg</td>
<td>1.45</td>
<td>7</td>
<td>0.125</td>
</tr>
<tr>
<td>Magnetic flux paths</td>
<td>2D</td>
<td>3D</td>
<td>2D</td>
</tr>
</tbody>
</table>

Among these materials listed in Table 2.1, AMM is available commercially in the form of very thin metal ribbon. In addition, it has much smaller thickness and is available in smaller widths compared to silicon steel which limits the core size. Furthermore, AMM is harder compared to silicon steel which makes it difficult to use conventional punching techniques. Due to the extremely thin (0.025mm) structure of AMM, it has iron losses which are typically an order of magnitude lower than conventional SI laminations (see Table 2.1). This makes AMM appealing for transformers and high-efficiency
variable-speed motor applications. However, AMM material also has lower saturation magnetic flux density and low stacking factor compared to SI, which reduces the achievable magnetic loading for a given volume.

On the other hand, SMC has the highest iron loss and the lowest permeability. Even though the iron loss of SMC is higher than SI at low frequencies, at high frequencies the iron losses (eddy-current losses specifically) are expected to be comparable due to the higher electric resistivity of SMC. In addition, SMC offers the potential for low-cost, high volume manufacturing and can readily produce complex 3-D core shapes which allows increased design flexibility.

2.1.1 Silicon Iron (SI)

Silicon iron is an iron alloy which is a good electrical conductor with typical 0-3% silicon added to increase resistivity which reduces eddy-current loss. The material has high saturation flux density, permeability and mechanical strength. It is manufactured in thin sheets by hot rolling iron into strips up to $1.3m$ wide in a rolling mill. To further minimise eddy-current loss, the laminations are usually coated to increase the electrical resistance between them. The laminations are also heat-treated in practice to reduce residual mechanical stresses and hence the hysteresis losses.

There are two main types of SI: grain-oriented and non-oriented.

- **Grain-oriented SI** is an anisotropic material which is specially processed to enhance the magnetic properties in the rolling direction. It has higher permeability, saturation magnetic flux density and lower core loss compared to non-oriented material. Its 1D flux path makes it suitable for transformers but generally unsuitable for rotating electric machines. In addition, it is more expensive than the non-oriented material.

- **Non-oriented SI** has a random crystal orientation and similar magnetic properties in both directions in the sheet. Hence it is widely used as laminated cores in electrical machines where the magnetic flux path has two dimensions. It is available in fully processed or semi-processed form and the standard thickness is between $0.12mm$ and $0.64mm$. It is less expensive to produce compared to the grain-oriented material and AMM.
In electrical machines, the stator and rotor are constructed by punching laminations and stacking them together (radial-field) or by winding the laminations into the final core shape and then cutting the slots (axial-field). Therefore, the use of laminations structure limits the magnetic flux paths to two dimensions.

### 2.1.2 Soft Magnetic Composite (SMC)

SMC consists of fine iron powder coated with a small amount of a binder which also serves to insulate the particles from one another. SMC can be moulded under high pressure to a wide range of shapes using a similar process to injection moulding of plastics. Then, the residual mechanical stress is reduced by heat treatment.

A key benefit of SMC is its ability to produce machines with 3D magnetic geometries and 3D flux paths. New machine design topologies with innovative shapes, reduced size and weight, and better winding utilisation are possible by using SMC’s 3D flux capability [7, 8]. SMC also offers a simplified manufacturing method at lower cost.

The insulation between the iron particles gives SMC a high electrical resistivity and hence low eddy-current loss. The moulding process however leaves high residual mechanical stresses in the iron particles and hence produces high hysteresis loss. Although SMC has higher iron loss compared to conventional silicon-iron (SI) at low frequencies where the hysteresis loss is dominant, the iron loss of SMC becomes comparable to SI at higher frequencies where the eddy-current loss becomes more important. This was experimentally demonstrated in [9].

The insulation between the iron particles in SMC also gives it a relatively low magnetic permeability \( \mu_R < 1000 \) which makes it better suited to permanent magnet (PM) machines than reluctance or induction machines.

**SMC in Rotating Machines**

The traditional laminated motor construction is inherently limited to 2D magnetic structures with 2D flux paths to avoid excessive eddy-current losses caused by flux passing perpendicularly through laminations. The 3D SMC core structures have been examined in the literature for:

- Transverse-field machines: for high torque, low speed, direct-drive applications [10].
• Tubular linear machines: for linear motion applications such as reciprocating compressors [11].

• Radial-field machines: laminated yoke with SMC teeth [9], segmented tooth construction [12, 13].

• Axial-field machines: high torque density applications such as small wind turbines [14–20].

2.1.3 Amorphous Magnetic Materials (AMM)

The research in amorphous magnetic materials (AMM) began in the 1960s [21]. Since then, it has attracted much attention from researchers regarding improving the technique of production, preserving its properties [22, 23] and also its applications in transformers [24, 25].

Amorphous iron is an iron alloy containing typical 6% silicon. It is produced commercially in thin ribbon form by rapid freezing of molten iron as it is poured on a fast spinning drum. The standard ribbon is 0.025-0.04 mm in thickness and up to 212 mm wide. As a result of the rapid solidification of the liquid iron, it has a non-crystalline structure with improved mechanical and electromagnetic properties compared to conventional steel alloys with a crystalline structure.

Amorphous iron has high permeability and very low iron loss especially at high frequencies. In AMM, the low eddy-current loss (due to the thin laminations) and low hysteresis loss (due to the non-crystalline structure) results in a very low total iron loss. This makes it a good candidate for high speed applications. However, amorphous iron has a lower operating flux density than conventional silicon iron due to a lower saturation flux density and low stacking factor (around 80%) due to thin laminations. Furthermore, the difficulty and high cost in material handling and processing of amorphous iron have limited its applications to simple shape core such as in transformers.

AMM in Rotating Machines

As stated above, the primary electric power application for AMM has been in ac power system distribution transformers. AMM transformers typically have only one-third of the core losses of conventional SI transformers [24, 25]. In addition, for transformers,
the AMM ribbon can be easily wound into a simple core structure and required limited cutting.

However, rotating machines generally has complex core shapes. Therefore, AMM should be cut accurately to achieve such shapes. It has been studied that conventional handling, punching and cutting methods can deform the cut surface and damage the electromagnetic properties of the materials. In addition, due to the hardness of AMM, it is difficult to punch the lamination which increases the cost of manufacturing.

Since the 1980s only a few attempts in utilising AMM in electric machines were reported in the literature, and the designs did not include complex cut shapes. In addition, poor performance of the early prototypes due to high losses and low output power has discouraged research efforts by industry. Therefore, AMM has not been utilised in commercial rotating electrical machines, mainly due to the absence of a commercially viable cutting technique. In recent years, however, there has been some improvements in AMM cutting, handling and core construction technology which has made the use of AMM cores in electrical machines much more practical [4, 9, 26, 27]. Furthermore, [28–49] are a list of granted US patents about constructing AMM machines. However, these methods are costly and not practical in terms of manufacturing repeatability and process cycle time. The following paragraphs summarise some of the previous rotating AMM-based machine studies.

General Electric Co. has been researching methods to utilise amorphous metal in rotating machines since the ribbon form became available. Some of the methods attempted include shaping the laminations when cooling the liquid amorphous metal alloy on the rotating cooling surface, punching, machining, etching or chemical milling, laser or electron beam cutting the individual yoke and teeth components, and bonding together to form the core. These techniques are covered in eight US patents [50–57]. Nevertheless, the methods are costly and not practical. Also, there were issues like deterioration of AMM magnetic properties due to the stress or cutting deformation during the construction process. In 1981, a laboratory AMM stator using the conventional approach of stacking laminations for a radial induction motor was constructed. Core loss tests were performed and the results were reported in [58, 59]. Reductions in core loss of up to 70-80% in amorphous iron machines compared to silicon iron machines were confirmed by the test results. However, two major deficiencies of amorphous metal cores were also identified: the low stacking factor increases the iron flux density and hence losses, and the magnetic properties deteriorate during processing.
In 1988, Boglietti et al. investigated utilisation of low iron loss magnetic materials such as amorphous and cubic texture materials based on finite element analysis [60]. In the simulations of the work, amorphous materials was found to be the best candidate as they offered the lowest iron loss especially at higher supply frequencies. However, cubic texture materials were used in constructing the prototype AC induction machines for validation testing due to the difficulty in assembling and cutting of the AMM ribbon to the desired shape. Therefore, the simulation results of the induction machine were not validated experimentally for AMM. A similar study was also reported in [61], AMM was not chosen due to the high cost of manufacturing complex core shapes using AMM.

The high iron loss in electrical machines using conventional magnetic materials under high speed operation has attracted attention of Japanese researchers. Fukao et al. reduced the core loss by utilising AMM [62] to increase the efficiency of a high-speed (48,000rpm) reluctance motor. In this work, the stator and rotor were constructed from 5cm width amorphous sheets by chemical etching methods. In addition, one of every five amorphous sheets in the stator core were coated with a dielectric. Significant core loss reductions of up to a factor of five was demonstrated experimentally. However, this was not a direct comparison of two similar size machines as the core loss for the silicon iron machines was calculated rather than measured.

In the search for simpler and cheaper manufacturing and design concepts that would maximised the utilisation of the AMM properties, an axial-field permanent magnet brushless dc motor was proposed by Jensen et al. in the 1990s [63]. In this research, a simple stator core was constructed from uncut tape-wound amorphous iron. The prototype machine was designed using a dual-rotor single-stator arrangement. The iron loss of the machine was not tested as the main focus was on developing the design equations and lumped-circuit model. The developed model in [63] indicated the possibility of high efficiency small size machines using AMM in axial-field PM machines.

Light Engineering Incorporated (LE) has been investigating and developing high performance variable-speed brushless PM motors and generators based on amorphous iron since 1996. LE has designed brushless AFPM machines with minimum and even without cutting, machining or milling the wound AMM toroid [28–41]. Some of the research work including the use of finite-element analysis software in design by LE were reported in [64–67]. LE has clearly demonstrated the feasibility of utilising AMM in rotating machines.
Another group developing AMM machines is AMM Technologies. One of their patents in [68] showed a cutting method based on high pressure liquids. The group claimed that they have developed precise and cost effective means in producing AMM machines [27]. The prototype machines of AMM Technologies demonstrated are showing 86% efficiency running from 3,000 to 24,000 rpm.

In recent years, more organisations such as Metglas Inc., AlliedSignal Inc., Honeywell International Inc., Amosense Co. Inc. and the Electric Power Research Institute have been granted patents in producing AMM machines [42–48]. The methods proposed in these patents construct individual teeth and yoke lamination segments and assemble the complete stator by adhesive bonding. The AMM segments were formed by photolithographic etching, high strain rate stamping, ”punch and die” or by folding and compressing. Alternatively, the slots in the AMM toroid were cut using electrochemical grinding. Also, a solid toroidal AMM core was proposed in [49] by crushing AMM ribbon into powder, mixing the other powder particles (for optimal uniform composition) and a binder before compressing with core mold and, by coating as in SMC cores. Nevertheless, the cost and performance of the machines were not reported in these patents.

2.2 AMM for AFPM Machines

The industry partner of this project, GMT has improved the AMM cutting technique based on using an abrasive water jet (AWJ). The AWJ approach produces no heat, offers distortion free cutting and does not harden the AMM. The cutting technique is patented in [26] which allows clean and accurate cutting of assembling blocks of laminated AMM. This is achieved by carefully angling the cutting head with respect to the workpiece during cutting. In addition, the technique is able to achieve cutting rates which are commercially acceptable, hence has potential for mass production of AMM electrical machines. Alternative cutting technologies such as laser cutting and chemical etching are not economic for mass production. Due to the commercial sensitivity, the specific details of the cutting technique are not included in this work, which is also used in the prototype machines in this thesis.

The tape-wound structure of an axial-field permanent magnet (AFPM) machine and its capability to handle inaccuracies in cutting make it well suited to utilise AMM. The slotted core structure is also suitable to be cut using the proposed AWJ cutting.
technique. Figure 2.1 shows the cut AMM AFPM stator prototypes produced using the technique developed by GMT.

![AMM-based axial-field motor stators cut by the GMT method.](image)

Figure 2.1. AMM-based axial-field motor stators cut by the GMT method.

Although the punched silicon-iron laminations radial-field machines are common, AFPM machines offer advantages such as short axial length, high output torque and high power density [2–4]. Therefore, the design of an AFPM machine using a cut AMM structure is considered and investigated in this research work.

### 2.2.1 Fractional-Slot Concentrated-Winding AMM Machines

Fractional-slot concentrated-winding permanent magnet machines have attracted strong interest due to their low cogging torque and copper loss, and suitability for fault-tolerant and flux weakening implementations. References [15, 69–85] examine the design, analysis and performance evaluation of such machines.

The slotted axial-field stator produced by cutting the AMM toroid (see Figure 2.1) is well suited for concentrated windings. This was demonstrated in [15] utilising SMC due to its simplicity in producing the 3D axial-field stator. Three identical concentrated-winding machines based on AMM, SMC and SI will be compared experimentally in Chapter 7.

Nevertheless, eddy-current losses in the rotor back-iron and surface-mounted permanent magnets for fractional-slot concentrated-winding designs can be very significant due to sub-harmonics from the fractional-slot windings [84–93]. The overall efficiency of the motor could be significantly reduced due to this effect. The eddy-current loss is also investigated in Chapter 8.
Chapter 2 Axial-Field Permanent Magnet Machines

2.3 Development of AFPM Machines

Axial-field PM machines were reported in 1830s, even before radial-field machines. However, since the introduction of the radial-field machines, they were widely accepted and became the mainstream configuration for electrical machines. The challenges with axial-field PM machines include the strong axial magnetic attraction force between the stator and rotor, the difficulty of construction, and the high manufacturing cost of the machine.

The developments in advanced power electronics and the recent substantial drop in the prices of rare-earth permanent magnets have given a strong boost in PM machine development. In addition, the increased awareness of the importance of energy conservation and environment factors have attracted interest in researching alternative electrical machines configurations offering higher power density, higher efficiency and greater cost-effectiveness.

2.3.1 Features and Advantages of AFPM Machines

PM machines are widely used in applications requiring high power density and high efficiency. For example, high performance PM machines are being developed for electric vehicles, wind generators, micro-hydro turbines, water pumps and household motor applications. The recent rise in popularity especially in automotive applications in axial-field permanent magnet (AFPM) machines is due to its short axial length, high torque, power density and high efficiency. In addition, these machines are suitable for ship propulsion drives, and in direct drive elevator applications offering small size, light weight, zero cogging and less operating noise.

The main differences of axial-field machines compared to conventional radial-field machines is in terms of the magnetic distribution and the direction of the flux flow. In axial-field machines, the flux flows axially in the direction parallel with the shaft of the machine while in radial-field machines it flow radially between the stator and rotor. The advantages and disadvantages of axial-field machines with respect to conventional radial-field machines are summarised in Table 2.2.
Table 2.2. Advantages and disadvantages of AFPM machines.

<table>
<thead>
<tr>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>• Short axial length</td>
<td>• Construction difficulties</td>
</tr>
<tr>
<td>• High torque and power density</td>
<td>- producing slotted stator core</td>
</tr>
<tr>
<td>• Tape-wound structure and its capability to handle inaccuracies in cutting (suitable for AMM)</td>
<td>- keeping uniform airgap with strong attractive force between stator and rotor</td>
</tr>
<tr>
<td>• Lower copper loss with concentrated windings</td>
<td>- complicated machine topology with multiple airgaps</td>
</tr>
<tr>
<td></td>
<td>• High windage losses at high-speed applications</td>
</tr>
</tbody>
</table>

### 2.3.2 Topologies and Geometries

Figure 2.2 gives the power ratings and operating speeds of some reported AFPM machines for a variety of applications, and the corresponding machine configurations are given in Table 2.3.

![Figure 2.2. AFPM machines power versus speed.](image)
AFPM machines have various topologies and configurations with different performance characteristics, construction methods and manufacturing costs. Table 2.4 gives a summary of the diverse topologies of AFPM machines including stator core and magnet structures and winding configurations.

AFPM machines can be designed in single-sided, double-sided and multi-disk configurations. Although the single-sided type (single stator and rotor) is the simplest construction, it is not common as the high magnetic attractive force between the rotor magnets and the stator can cause bearing problems. The attractive force can be
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Table 2.4. AFPM machines topologies and geometries.

<table>
<thead>
<tr>
<th>Topology</th>
<th>Stator core</th>
<th>Stator Winding</th>
<th>Rotor</th>
</tr>
</thead>
<tbody>
<tr>
<td>single-sided</td>
<td>slotted</td>
<td>distributed</td>
<td>surface PM</td>
</tr>
<tr>
<td>double-sided</td>
<td>slotless</td>
<td>concentrated</td>
<td>interior PM</td>
</tr>
<tr>
<td>multi-stage</td>
<td>coreless</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

eliminated by counterbalancing it with two stators or rotors in the double-sided configuration. The double-sided configuration offers twice the output torque and can be designed to further reduce copper and iron losses. Most published AFPM machines are of this type. In order to obtain higher output torque, multi double-sided machines can be placed on the same shaft to form the multi-stage type topology [104]. In general, the choice of configurations also depends strongly on the application and operating environment.

The stator in AFPM machines can be slotted, slotless or coreless. In a slotted design, the airgap is small allowing higher magnetic loadings and thinner magnets to be used. Nevertheless, the stator teeth in slotted designs have iron losses, saturate at high flux densities and produce cogging torque. These can be eliminated in a slotless design which uses a simple toroidal stator core and require no cutting of the stator slots. However, the slotless design require a large airgap hence thicker PMs. In addition, winding inductance is lower in such topology requiring a higher inverter switching frequency, poorer cooling of the stator winding and higher eddy-current losses in the conductors as they are directly exposed to the rotor magnetic field.

As shown in Table 2.4, there are two types of magnet configuration in AFPM machines: the surface-mounted and interior (or buried) permanent magnet. The surface-mounted machines offer simpler control and analysis (non-salient design), high peak torque capability (low stator leakage inductance) and large effective airgap with low winding inductance. However, they are not suitable to operate above the base speed, and their mechanical integrity is lower under high speed operation. For interior permanent magnet machines, the permanent magnets are embedded in the rotor iron [105–108]. This type is suitable for high-speed operation as the magnets are retained by the rotor iron. It also offers low magnet eddy-current losses, low demagnetisation risk,
lower controller switching frequency and improved field-weakening perform. However, interior PM machine design and construction is more complicated compared to the surface-mounted rotor types.

The stator windings for slotted machines can be generalised into two types that are the distributed and concentrated windings. For distributed windings, the coils span more than one slot. For concentrated windings, the coils are wound around a single tooth only. They can be wound around each tooth (double-layer concentrated winding) or alternate teeth (single-layer concentrated winding). In general, concentrated windings offers lower copper loss due to their shorter end windings, simpler winding automation and greater magnetic and electrical isolation between phases, but generally has smaller winding factors and larger winding space harmonics compared to distributed windings. The shorter end windings can result in shorter total length and lower manufacturing cost. Concentrated windings also provide higher inductance compared to distributed windings for the same magnet flux-linkage, hence are a good candidate for flux-weakening designs [109]. In double-sided machines, the stator windings can be connected either in parallel or in series. The series connection is generally preferred as it avoids circulating currents. However, a parallel connection allows the motor to continue to operate even if one stator winding is open-circuited.

2.3.3 PM Motor Drives for AFPM

In permanent magnet machines, torque is produced by the rotor permanent magnets wanting to align with the rotating magnetic field from the stator winding when it is excited by square-wave currents in trapezoidal back-EMF (also known as brushless DC machines) and sine-wave currents in sinusoidal back-EMF (also known as AC machines) brushless PM machines [110, 111].

In brushless DC motors, the phase back-EMF waveform is ideally trapezoidal in shape and only two phases are conducting at any given time. Based on the rotor position, the appropriate stator winding phases are energised to produce maximum torque. In a sinusoidal AC motor, the motor operates on the principle of a smoothly rotating magnetic field with sine-wave current excitation fed into the windings. Hence, the phase back-EMF waveform is ideally sinusoidal and all three phase windings conduct current at any given time.
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Axial-Field Permanent Magnet Machines

It was shown in [112] using analytical comparisons that there is no significant torque density difference between sine-wave and square-wave machines. The main differences between the two current excitation modes are listed in Table 2.5. The fast dynamic response and smooth output torque (including at high speed) produced by sine-wave machines make them suitable for precision machine tools. However, they require a precision rotor position sensors to generate a sinusoidal reference current, which result in more complex control and higher cost compared to the square-wave drives. Square-wave machines are usually applied in low-cost variable-speed applications where the smooth output torque is not required.

In this research, square-wave motor drive with a slotted-stator machine was implemented and pulse-width modulation (PWM) was used to vary the amplitude of the six-step inverter square-wave output voltage to control the current and hence steady-state speed of the motor (see Sections 4.2.3 and 7.2.3).

<table>
<thead>
<tr>
<th>Sinusoidal AC</th>
<th>Trapezoidal DC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Best dynamic performance</td>
<td>Low-cost variable-speed applications</td>
</tr>
<tr>
<td>(fast response, smooth torque)</td>
<td>(slower response, high torque ripple)</td>
</tr>
<tr>
<td>Complex control</td>
<td>Simple control</td>
</tr>
<tr>
<td>Accurate rotor position sensor</td>
<td>Simple rotor position sensor</td>
</tr>
<tr>
<td>(encoder such as resolver)</td>
<td>(Hall-Effect)</td>
</tr>
<tr>
<td>High cost</td>
<td>Lower cost</td>
</tr>
</tbody>
</table>

Table 2.5. Sinusoidal AC and trapezoidal DC PM motors comparison.

2.4 Fabrication of AFPM Stator Cores

2.4.1 Laminated SI Stator Core

Figure 2.3 shows three alternative methods for fabricating slotted AFPM stator cores from silicon-iron laminations. The simplest method is to wind the laminations into a cylindrical coil and then machine the radial stator slots. Alternatively, the slots can be punched in the steel strip at a carefully calculated spacing and then wound to form the slotted cylindrical core [111]. The third technique is proposed in [113] and consists of
constructing the stator core using individual trapezoidal tooth segments. The lamination strip is first folded with the help of the grooves in a zigzag form with each segment proportional to the radius of the slot. The zigzag laminated segments are then compressed and secured to form the tooth segments (see Figure 2.4). The complete stator is then assembled by adhesive bonding the teeth and yoke lamination segments.

Figure 2.3. Slotted AFPM stator assembly production flow chart for SI showing three alternative construction methods.

Figure 2.4. SI stator assembly method [111,113].
2.4.2 Soft Magnetic Composite Stator Core

Figure 2.5 shows the AFPM stator fabrication flow chart based on SMC. SMC can be moulded to a wide range of 3D shapes using a die. For prototypes, the stator is usually constructed by machining a solid block of SMC as shown in Figure 2.6. Note, the slot packing factor is improved by using rounded edges on the stator teeth (see Figure 2.6(c)) hence allowing the use of thinner insulation between the winding and the iron.

![Diagram](image)

**Figure 2.5.** Slotted AFPM stator assembly production flow chart of SMC.

![Stator Images](image)

(a) 32mm stator  
(b) 110mm stator  
(c) 110mm stator

**Figure 2.6.** Example SMC axial-field motor stators.
2.4.3 AMM Stator Core

Due to the cutting and handling difficulties of AMM, there has been only limited research done utilising AMM in rotating electrical machines. Figure 2.7 shows the proposed AFPM stator fabrication flow chart utilising AMM for slotless machines and for slotted machines using the proposed cutting approach.

![AMM Stator Core Diagram]

**Figure 2.7.** Stator assembly production flow chart for slotless and slotted AFPM AMM cores.

2.5 AFPM Rotor

The magnetic flux in AFPM machines is produced by permanent magnets rather than a field winding. Hence, the rotor consists of permanent magnets and back-iron. The back-iron is added to increase the flux density and to reduce flux leakage. High energy density permanent magnets such as rare-earth (Neodymium) magnets are often utilised in AFPM machines. The magnetic field produced by the rotor induces the back-EMF voltage in the stator windings. The amplitude and shape of the back-EMF waveforms are affected by the magnet type, magnet spacing and the presence of rotor back-iron. This is investigated in this thesis using finite-element analysis and experimental tests (see Sections 4.3.2 and 7.4.3).
The surface-mounted rotor configuration is used in this project for simplicity. The rotor is constructed by gluing the magnets on the mild steel back-iron using epoxy. The eddy-current losses in the magnets and back-iron can be significant especially for fractional-slot stator winding machines under high speed operation. In order to reduce the eddy-current losses, laminated rotor back-iron, segmented magnets and bonded magnets can be used.

Two types of neodymium (NdFeB) magnets are commonly used in PM machines: sintered and bonded types. Sintered magnets are anisotropic and have a maximum energy product in the range between 80 to 350 $kJ/m^3$. On the other hand, bonded magnets are isotropic with a maximum energy product up to 65 $kJ/m^3$. The typical properties of these magnets are given in Table 2.6. The choice of PM material depends on the design and application.

<table>
<thead>
<tr>
<th>Characteristics</th>
<th>Sintered</th>
<th>Bonded</th>
</tr>
</thead>
<tbody>
<tr>
<td>Residual Induction $B_r$ [T]</td>
<td>0.7-1.4</td>
<td>0.45-0.65</td>
</tr>
<tr>
<td>Coercive Force $H_c$ [kA/m]</td>
<td>450-1,600</td>
<td>310-420</td>
</tr>
<tr>
<td>Energy Product $(BH)_{max}$ [kJ/m$^3$]</td>
<td>80-350</td>
<td>35-65</td>
</tr>
<tr>
<td>Recoil Permeability $\mu_{rec}$ [-]</td>
<td>1.04-1.3</td>
<td>1.15</td>
</tr>
<tr>
<td>Temperature coefficient of $B_r$ to 100°C [%/°C]</td>
<td>-(0.07-0.16)</td>
<td>-0.2</td>
</tr>
<tr>
<td>Temperature coefficient of $H_c$ to 100°C [%/°C]</td>
<td>-(0.37-0.9)</td>
<td>-0.4</td>
</tr>
<tr>
<td>Resistivity [$\mu\Omega m$]</td>
<td>1.4-1.6</td>
<td>180</td>
</tr>
<tr>
<td>Density [$kg/m^3$]</td>
<td>7,200-7,500</td>
<td>5,000-6,000</td>
</tr>
</tbody>
</table>

The high residual induction sintered magnet would produce a higher magnetic loading compared to the lower strength bonded magnet and hence result in a higher torque density, lower copper loss and less magnet mass. On the other hand, the high electric resistivity of the bonded magnet plays an important role in minimising the magnet eddy-current loss that can be significant (see Section 7.4.3). In addition, the operating temperature should be taken into consideration as both the residual induction and intrinsic coercivity are temperature dependent.
2.6 Conclusions

Some recently developed prototype concentrated-winding AFPM machines are given in [14, 71, 94–96, 103]. Their core materials, configurations, power ratings, operating speeds and measured efficiencies are summarised in Table 2.7. All the machines were constructed with sintered neodymium magnets. As shown in the table, the highest reported efficiency was achieved with a double-sided configuration (93%) while for the single-sided configuration this was lower (80%). A non-optimised SMC based machine only achieved 50% efficiency.

<table>
<thead>
<tr>
<th>Table 2.7. Efficiency of AFPM prototypes.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Machine</td>
</tr>
<tr>
<td>Stator material</td>
</tr>
<tr>
<td>Number of stator teeth</td>
</tr>
<tr>
<td>Configuration</td>
</tr>
<tr>
<td>Number of poles</td>
</tr>
<tr>
<td>Outer diameter (mm)</td>
</tr>
<tr>
<td>Airgap length (mm)</td>
</tr>
<tr>
<td>Speed (rpm)</td>
</tr>
<tr>
<td>Power (W)</td>
</tr>
<tr>
<td>Efficiency (%)</td>
</tr>
</tbody>
</table>

In this work, the use of AMM in producing high-efficiency, single-sided AFPM machines was investigated. Extensive experimental results are reported in Chapters 4, 7 and 8.

The use of alternative magnetic materials in electric machine design offers the opportunity to achieve lower losses specifically at higher operating frequencies. The properties of two emerging materials AMM and SMC were presented and compared with the conventional SI. Also, the existing utilisation of AMM and SMC in rotating electrical machines was described.
The benefits and current limitations of applying AMM in rotating machines were discussed. The background of AFPM machines is introduced including its features, topologies, fabrication and materials. The utilisation of AMM in AFPM machines was proposed. The AFPM machine was chosen to utilise AMM in this work due to its tape-wound structure and capability to handle inaccuracies in cutting. In addition, the AFPM machine configuration can provide the additional benefits such as high torque and power density.

Therefore, the design of an AFPM machine using a cut AMM structure is considered and investigated in this work. The work aims to provide a better understanding of the design approach to utilise AMM for high-efficiency electrical machines.
In this chapter, a comprehensive investigation was conducted to examine the iron loss characteristics of the AMM non-slotted and slotted cores at various frequencies. This includes detailed 3D finite-element modelling. Two compensation methods for the non-uniform flux density in the slotted cores were introduced to estimate the iron loss characteristics. In addition, analytical modelling was utilised to analyse the increased loss due to interlamination eddy-current loss.

In order to quantify and compare the magnitude of the iron losses, similar tests were also performed on SI and SMC cores.
3.1 Iron Loss Measurement Method

To quantify the magnitude of the iron losses at higher frequencies for the three magnetic materials (SI, SMC and AMM, see Table 2.1), a magnetic flux density versus field strength (BH) curve is utilised. A similar measurement method as in [109, 114] was conducted at different frequencies.

Figure 3.1(a) illustrates the principle of the experimental setup used to measure the BH-curve of a uniform toroidal magnetic core. In the figure $N_{\text{main}}$ and $N_{\text{sense}}$ are the number of turns of the main and the sense windings, $l_{\text{core}}$ is the mean length of the magnetic flux path, $A_{\text{core}}$ is the cross-sectional area of the magnetic core, and $i_{\text{main}}$ and $v_{\text{sense}}$ are the instantaneous values of the main winding current and sense winding induced voltage respectively.

![Measurement setup](image)

![Equipment setup](image)

Figure 3.1. Iron loss test measurement and equipment setup.

To prepare the device under test (DUT), firstly the core was covered with insulating tape to protect the wire from being cut by the sharp edges of the non-slotted core. Secondly, two sets of windings were wound on the core: the main winding and the sense winding (see Figure 3.2(a)). During the tests, the main winding was excited by a 50Hz variable AC voltage source (via an isolating step-down transformer and an auto-transformer) or a variable-frequency signal generator which was connected through an amplifier (Kepco BOP 50-8M) as shown in Figure 3.1(b). Then the main winding current and the sense winding voltage waveforms were measured via high bandwidth current and voltage transducers (Hameg HZ56 current probe and Yokogawa differential voltage probe model 700925) whose outputs were recorded using a digital oscilloscope (Tektronix TDS 340A). A power analyser (Voltech PM3000ACE) was also used to provide a more accurate input power measurement.
In order to obtain the iron loss for the DUT, the measured data was processed using a custom-written spreadsheet program in Excel. The core flux density waveform $B(t)$ was obtained by integrating the search coil voltage $v_{\text{sense}}(t)$, which is given by Equation 3.1 where $A_{\text{core}}$ takes into account the stacking factor from Equation 3.4.

$$B(t) = \frac{1}{N_{\text{sense}}A_{\text{core}}} \int v_{\text{sense}}(t)dt$$  \hspace{1cm} (3.1)

The magnetic field strength waveform $H(t)$ was obtained using the main coil current $i_{\text{main}}(t)$, and is given by

$$H(t) = \frac{N_{\text{main}}i_{\text{main}}(t)}{l_{\text{core}}}$$  \hspace{1cm} (3.2)

The core loss of the DUT was obtained by averaging the product of the search coil voltage and the main coil current, and is given by

$$P_{fe} = \frac{N_{\text{main}}}{N_{\text{sense}}} \frac{1}{T} \int_0^T v_{\text{sense}}(t)i_{\text{main}}(t)dt$$  \hspace{1cm} (3.3)

The core loss is divided by the mass of the test core to obtain the iron loss in Watts/kg. It should be noted that the iron loss can also be obtained using the BH-curve, as the iron loss is directly proportional to the area enclosed by the BH-curve.

The above test procedure was repeated for different frequencies and current limited by the ratings of the power supplies and main winding. Figure 3.3 shows a comparison of the measured BH-curves for SI, SMC and AMM. AMM has the lowest iron loss and
hence the smallest loop area. On the other hand, SMC showed a large area indicating high iron loss.

Figure 3.3. Measured BH-curves of SI, SMC and AMM at 50Hz and B=1T.

The iron loss measurements conducted in [23, 114–116] demonstrated the low iron loss characteristics of AMM at 50Hz and in [9] at high frequencies. In addition, [115, 116] also showed that pulse-width modulation (PWM) inverter harmonics have the effect of increasing the eddy-current loss and hence iron loss in the AMM core. However, these are not investigated here as it is not in the scope of research in this thesis.

Figure 3.4 gives the measured $v_{\text{sense}}$ and $i_{\text{main}}$, calculated $B$ (Equation 3.1), $H$ (Equation 3.2) and $P_{fe}$ (Equation 3.3) waveforms at unsaturated (0.25A) and saturated current (2.5A) values for the grain-oriented SI non-slotted core. It can be seen in the results that at low current values, the waveforms showed a sinusoidal behavior. On the other hand, the waveforms were distorted especially the measured current waveform, as saturation occurred in the core.

Figure 3.5 gives the measured BH-curves of the grain-oriented SI core obtained by plotting $B$ as a function of $H$. The measured peak flux density and field strength points at various currents were plotted in Figure 3.6(a) and compared to the values given in the data sheet [117]. The values showed good correspondence and hence verified the experimental results. In addition, the measured iron loss density at various flux densities were compared with the values from the data sheet in Figure 3.6(b). Again, the values matched closely.
Figure 3.4. Measured voltage $v_{\text{sense}}$ and current $i_{\text{main}}$, power $P_f$, flux density $B$ and field strength $H$ waveforms at 0.25A (left, unsaturated) and 2.5A (right, saturated) currents.
Chapter 3  AMM Iron Loss Measurements

Figure 3.5. Measured BH-curve of grain-oriented SI at 50Hz and two different current levels: 0.25A and 2.50A.

Figure 3.6. Comparison of measured values (Meas) and data sheet (Data) BH-curves and iron loss curves of grain-oriented SI at 50Hz.

3.2 Non-Slotted Core Iron Loss Tests

The five test samples used are shown in Table 3.1. These are a solid SMC (Figure 3.2(a)) core, stacked non-slotted ring shaped 0.5mm laminations of non-oriented SI, tape-wound 0.27mm grain-oriented SI core (Figure 3.2(b)), tape-wound 0.025mm uncoated AMM core (Figure 3.2(c)) and tape-wound 0.020mm coated AMM (AMM2) core. Table 3.1 summarises the specifications of the test cores. The stacking factor of the laminated toroidal cores in this table was estimated based on Equation 3.4 which is the...
fraction of volume in the core structure that is occupied by iron. Note that, this stacking factor is accounted in the flux density values presented in this thesis.

\[
\text{stacking factor} = \left( \frac{M}{\rho_M \cdot \text{Vol}_{cal}} \right) \times 100\% \quad (3.4)
\]

where \(M\) is the mass, \(\rho_M\) is the material density and \(\text{Vol}_{cal}\) is the calculated core volume.

**Table 3.1. Five cores in the tests: dimensions and some of the specifications.**

<table>
<thead>
<tr>
<th>Material</th>
<th>Inner Diameter (mm)</th>
<th>Outer Diameter (mm)</th>
<th>Height (mm)</th>
<th>Mass (kg)</th>
<th>Estimated stacking factor (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SMC: Somaloy 550+0.6%LB1</td>
<td>55</td>
<td>110</td>
<td>30</td>
<td>1.29</td>
<td>solid</td>
</tr>
<tr>
<td>Non-oriented SI: Lycore 230</td>
<td>38</td>
<td>50</td>
<td>10</td>
<td>0.07</td>
<td>20 stacked laminations</td>
</tr>
<tr>
<td>Grain-oriented SI: 27RGH100</td>
<td>45</td>
<td>110</td>
<td>30</td>
<td>1.77</td>
<td>98</td>
</tr>
<tr>
<td>Uncoated AMM: Metglas 2605SA1</td>
<td>45</td>
<td>110</td>
<td>30</td>
<td>1.47</td>
<td>90</td>
</tr>
<tr>
<td>Coated AMM: Vitroperm 800</td>
<td>80</td>
<td>100</td>
<td>25</td>
<td>0.34</td>
<td>66</td>
</tr>
</tbody>
</table>

Due to limited material resources, uncoated 2605SA1 AMM ribbon was utilised to construct the tape-wound AMM ribbon core (AMM1, see Figure 3.2(c)). As it can be seen, absence of insulation between laminations results in higher iron loss due to the additional eddy-current loss. In addition, the high tension used in winding the lamination core to obtain a high stacking factor results in higher iron loss due to high tensile stress on AMM, which also increases the electrical contact between the layers. This was discussed in [114], where the authors suspect that the discrepancy in the magnetic characteristics of the measured AMM materials is caused by mechanical stress in the wound core. This was further verified in [23] which showed high core loss in uncoated core wound with high tension compared to coated cores. As listed in Table 3.1, a small toroid of coated AMM ribbon (AMM2) was also tested in this research.
3.2.1 Measurement Results at 50Hz

The procedure explained in Section 3.1 was used to test the five cores mentioned above using a 50Hz sinusoidal excitation. The measured saturation flux density (see Table 3.2) and iron loss (see Figure 3.7) are compared to their manufacturer’s data sheet values where available [117–120].

<table>
<thead>
<tr>
<th>Material</th>
<th>Flux Density B @ respective H A/m</th>
<th>Data sheet (T)</th>
<th>Test (T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SMC: Somaloy</td>
<td>1.40</td>
<td>1.10</td>
<td></td>
</tr>
<tr>
<td>550+0.6%LB1</td>
<td>@ 7,400A/m</td>
<td>@ 7,400A/m</td>
<td></td>
</tr>
<tr>
<td>Non-oriented SI: Lycore 230</td>
<td>-</td>
<td>1.65</td>
<td></td>
</tr>
<tr>
<td>Grain-oriented SI: 27RGH100</td>
<td>1.80</td>
<td>1.77</td>
<td></td>
</tr>
<tr>
<td></td>
<td>@ 95A/m</td>
<td>@ 95A/m</td>
<td></td>
</tr>
<tr>
<td>Uncoated AMM: Metglas 2605SA1</td>
<td>1.57</td>
<td>1.30</td>
<td></td>
</tr>
<tr>
<td></td>
<td>saturation</td>
<td>@ 1,300A/m</td>
<td></td>
</tr>
<tr>
<td>Coated AMM: Vitroperm 800</td>
<td>1.21</td>
<td>1.28</td>
<td></td>
</tr>
<tr>
<td></td>
<td>saturation</td>
<td>@ 77A/m</td>
<td></td>
</tr>
</tbody>
</table>

The iron loss curves in Figure 3.7 were curve fitted using Equation 3.5. Table 3.3 compares the value of the two iron loss parameters in the equation for the test cores with the data sheet values where available.

\[
P_{feDen} = KB^α
\]

where \( K \) is the iron loss at a flux density of 1T and a frequency of 50Hz, and \( α \) is the flux density term which represents the iron loss dependence on flux density and is the slope of the curves in Figure 3.7.

From Table 3.2, Figure 3.7 and Table 3.3, it can be seen that the grain-oriented SI measurements match the data sheet. Also comparing the loss at 1T and 50Hz, the non-oriented SI exhibits about 10 times higher iron loss compared to the grain-oriented SI. The coated AMM (0.01W/kg) has the lowest measured iron loss followed by grain-oriented SI (0.3W/kg), uncoated AMM (0.45W/kg), non-oriented SI (3W/kg) and SMC.
Figure 3.7. Comparison of measured and data sheet iron loss versus peak flux density curves at 50Hz: a) linear flux density scale, b) logarithm flux density scale. Material used in measurements, Table 3.1: non-oriented SI (SI Meas), grain-oriented SI (SIO Meas), SMC (SMC Meas), uncoated AMM (AMM1 Meas) and coated AMM (AMM2 Meas). Values from data sheets provided by the manufacturers: grain-oriented SI (SIO Data), SMC (SMC Data) and uncoated AMM (AMM1 Data).
Table 3.3. Iron loss and flux density term ($\alpha$ value) of 50Hz sinusoidal test results.

<table>
<thead>
<tr>
<th>Material</th>
<th>Loss @ B = 1T (W/kg)</th>
<th>Flux Density B Term ($\alpha$ value)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Data Sheet</td>
<td>Test</td>
</tr>
<tr>
<td>SMC: Somaloy 550+0.6%LB1</td>
<td>7.8 (SMC Data)</td>
<td>22 (SMC Meas)</td>
</tr>
<tr>
<td>Non-oriented SI: Lycore 230</td>
<td>-</td>
<td>3.0 (SI Meas)</td>
</tr>
<tr>
<td>Grain-oriented SI: 27RGH100</td>
<td>0.32 (SIO Data)</td>
<td>0.30 (SIO Meas)</td>
</tr>
<tr>
<td>Uncoated AMM: Metglas 2605SA1</td>
<td>0.065 (AMM Data)</td>
<td>0.55 (AMM1 Meas)</td>
</tr>
<tr>
<td>Coated AMM: Vitroperm 800</td>
<td>-</td>
<td>0.011 (AMM2 Meas)</td>
</tr>
</tbody>
</table>

The coated AMM results show about 30 times lower iron loss compared to the grain-oriented SI core which is consistent with the results given in [114]. Furthermore, the measured peak flux density was 1.28T which is about 23% lower than given in the data sheet of 2605SA1.

On the other hand, the measured loss of uncoated AMM was about 80% higher than the grain-oriented SI core and eight times greater than the data sheet values. This was expected as uncoated AMM was wound with high tension and was not insulated. In addition, Table 3.3 shows that the measured peak flux density is 20% lower compared to the data sheet and the experimental results given in [23, 114].

For SMC, the measured iron loss value is found about 2.8 times higher than the loss given in the data sheet of Somaloy 550+0.6%LB1. It is possible that this increase is due to the different heat treatment on the test sample. In addition, the measured peak flux density is found 27% lower than the data sheet values (see Table 3.3).

In regards to the flux density term $\alpha$ (which describes the flux density dependence of iron loss) the grain-oriented SI and uncoated AMM cores shows comparable measured and data sheet values. For the SMC, there is a 13% difference between the measured and data sheet values which could also be due to the different end heat treatment of the test core. Comparing the $\alpha$ value of the test cores, coated AMM shows the highest
value followed by the grain-oriented SI and SMC. On the other hand, the non-oriented SI and uncoated AMM have the lowest $\alpha$ values.

As predicted, AMM (AMM2 in Figure 3.7) offers extremely low iron loss compared to SI and SMC. Although SMC has higher iron losses than non-oriented SI at 50 Hz, the predicted iron loss of SMC is comparable to non-oriented SI at higher frequencies. This is due to the fact that eddy-current loss becomes dominant at the high frequency region and the high electrical resistivity in SMC leads to lower eddy-current loss compared to SI. This was experimentally demonstrated in [9].

### 3.2.2 Measurement Results in 50 Hz to 1,000 Hz Range

In order to describe the variation of iron loss with frequency, a frequency term was included into Equation 3.5 and is given below in Equation 3.6. The frequencies were normalised based on the 50 Hz case. The parameter $\beta$ is the frequency term which represents the variation of iron loss with frequency. Table 3.4 provides a list of the constant, the flux density and the frequency terms which are calculated based on the measured results of the test cores over a range of frequencies from 50 Hz to 1,000 Hz. The term $\beta$ was also calculated from the data sheet iron loss curves and included where available. Figures 3.8 to 3.11 are provided to illustrate the iron loss plots on a logarithmic scale for the materials, which utilise the parameter $\beta$.

$$P_{feDen} = KB\alpha\left(\frac{f}{50}\right)^\beta$$  \hspace{1cm} (3.6)

Figure 3.8 provides a comparison of the measured iron loss for the uncoated AMM1 lamination non-slotted core with the values from the data sheet for 2605SA1 (AMM) ribbon at 50 Hz, 100 Hz, 200 Hz, 500 Hz and 1,000 Hz. It can be seen that the measured AMM1 has about eight times higher iron loss compared to the data sheet at 1 T, 50 Hz due to the reasons discussed previously. Nevertheless, the magnitude difference reduces as frequency increases and is down to about four times larger than the data sheet values at 1,000 Hz. This is reflected in the lower measured frequency term compared to the data sheet value. The plots in Figure 3.8 showed comparable slope and hence similar flux density terms as in Table 3.4.
Table 3.4. Comparison of data sheet versus measured values for constant, flux density and frequency terms of the iron loss equation.

<table>
<thead>
<tr>
<th>Material</th>
<th>Constant Term (K value)</th>
<th>Flux Density Term (α value)</th>
<th>Frequency Term (β value)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Data Sheet</td>
<td>Test</td>
<td>Data Sheet</td>
</tr>
<tr>
<td>SMC: Somaloy 550+0.6%LB1</td>
<td>8.12</td>
<td>25.4</td>
<td>1.64</td>
</tr>
<tr>
<td>Non-oriented SI: Lycore 230</td>
<td>-</td>
<td>2.50</td>
<td>-</td>
</tr>
<tr>
<td>Grain-oriented SI: 27RGH100</td>
<td>0.32</td>
<td>0.30</td>
<td>1.99</td>
</tr>
<tr>
<td>Uncoated AMM: Metglas 2605SA1</td>
<td>0.059</td>
<td>0.46</td>
<td>1.75</td>
</tr>
<tr>
<td>Coated AMM: Vitroperm 800</td>
<td>-</td>
<td>0.011</td>
<td>-</td>
</tr>
</tbody>
</table>

Figure 3.8. Comparison of measured versus 2605SA1 data sheet (AMM) iron loss for uncoated AMM (AMM1) non-slotted core over 50Hz to 1,000Hz range.
Chapter 3

AMM Iron Loss Measurements

Figure 3.9. Comparison of measured versus 2605SA1 data sheet (AMM) iron loss for coated AMM (AMM2) non-slotted core over 50 Hz to 1,000 Hz range.

Since the data sheet values of the coated AMM was not available, the measured results of coated AMM was compared with the data sheet values of another AMM 2605SA1. The measured iron loss was six times lower than the data sheet value at 1 T, 50Hz, which increased significantly with frequency (about 16 times lower at 1,000 Hz). This is associated with a lower frequency term (dependence) for the measured results versus the data sheet value. The coated AMM2 showed a steeper slope compared with the data sheet of 2605SA1 ribbon (see Figure 3.9 and flux density term in Table 3.4). This is associated with a 20% larger flux density term. Hence, from the above results it can be concluded that AMM2 is of different ribbon than 2605SA1 (AMM). The Metglas 2714a has the lowest iron loss of 0.03 W/kg at 60Hz and B = 1T available in [119]. This result is encouraging as it means that the iron loss of the AMM machine prototypes constructed in this work can be further improved with coated AMM.

Comparing the uncoated AMM1 to coated AMM2 test results, the iron loss is about 50 times higher at 50Hz and up to 150 times higher at 1,000 Hz due to larger frequency term (dependence). On the other hand, the flux density term was about 20% smaller indicating that the uncoated AMM1 is less sensitive to flux density changes.

The measured iron loss for grain-oriented SI (SIOMeas) and SMC (SMCMMeas) non-slotted cores were compared to the manufacturer’s data sheet value (SIOData, SMC-Data) over a similar frequency range in Figures 3.10(a) and 3.11(a). The grain-oriented
SI measurements matched closely at all tested frequencies with the data sheet curves and iron loss terms. Also, all the terms matched closely with the values from data sheet (see Table 3.4). Hence, it was concluded that the measurement method was accurate.

![Graph of铁芯损耗测量](image)

**Figure 3.10.** Comparison of measured versus data sheet iron loss for SIO and AMM1 non-slotted core over 50\(Hz\) to 1,000\(Hz\) range
For SMC, Figure 3.11(a) and Table 3.4 showed a three times higher iron loss at 50Hz compared to the data sheet and up to 10 times at 1,000Hz which indicates significant discrepancies in both the constant and frequency terms. The increase in iron loss could be caused by different heat treatment. The slope and hence flux density term was some what larger for the measured results than the data sheet.

Figure 3.11. Comparison of measured versus data sheet iron loss for SMC and AMM1 non-slotted core at 50Hz over 1,000Hz range.
In addition, the measured iron loss of grain-oriented SIOMeas and the SMCMeas nonslotted cores were compared with the uncoated AMM1Meas cores over a similar frequency range. These materials were later utilised to construct three identical machines for comparison in Chapter 7. These results were also given in Figures 3.10(b) and 3.11(b). At 50 and 100 Hz, grain-oriented SI showed lower iron loss of about 2 times in magnitude. Similar iron loss was observed at 200 Hz for the two materials but at high frequency regions the loss increased much faster for SI and the iron loss ratio was up to 1.5 times at 1,000 Hz. As for SMC, significant higher iron loss was observed compared to the uncoated AMM1. The iron loss of SMC at 50 Hz was about 80 times higher than AMM1 and about 200 times higher at 1,000 Hz.

Comparing the iron loss terms calculated from the test results of the above materials from Table 3.4, SMC has the highest constant and frequency terms with about 50 times and 1.6 times higher respectively compared to uncoated AMM. The flux density terms of all the materials were within 10% of the ideal value of two. Therefore, highest iron loss was expected for SMC. Although the grain-oriented SI has the lowest constant term, it also showed higher flux density (14%) and frequency (43%) terms compared to the uncoated AMM. As a result, the iron loss of uncoated AMM would be lower than the grain-oriented SI at higher flux density and frequency regions. This is consistent with the measured results plotted in Figure 3.10(b).

### 3.3 Slotted Core Iron Loss Tests

#### 3.3.1 3D Finite Element Modelling

The iron loss data of the various materials obtained above using the non-slotted core measurement were utilised in the finite-element analysis (FEA) software package, JMAG. In JMAG, the iron loss of the machine was simulated based on the flux density in each element in the model, and the iron loss characteristics of the material measured in the previous section was utilised as will be described in Chapter 6.

Finite-element models of the experimental cores were constructed in JMAG and simulated using the experimental iron loss curves (AMM1) to examine the accuracy of the simulated iron loss. The coils were modeled covering the outer surface of the core as shown in Figure 3.12. Note that the upper small section of the energising windings was
not shown in the figure so that the core can be seen clearly. The secondary sense winding was not required to be modelled in JMAG as the flux can be calculated directly.

![Figure 3.12](image)

**Figure 3.12.** FEA iron loss model (quarter): non-slotted and slotted cores with the top section of the FE coil model removed for clarity.

The 3D FE model is able to simulate the non-uniform flux distribution in the slotted core. Examples of the simulated magnetic flux density using vector plots are shown in Figure 3.13. With the non-slotted core, the flux density is higher at the inner diameter as expected. With the slotted case, Figure 3.13(b) indicates that the flux density is highest at the inner diameter of the teeth. In both cases, the flux density is obtained by dividing the circumferential flux by the yoke area.

![Figure 3.13](image)

**Figure 3.13.** FEA non-slotted and slotted models magnetic flux density vector view.

Figure 3.14(a) shows the comparison between the measured and FEA simulated iron loss for the non-slotted core, which closely matches the test results. Then, the iron loss simulated with the slotted core model and the measured results are given in Figure
3.14(b). Note that, same flux density estimation approach used for the measured values ("AMM3Meas") was performed on the simulated values for consistence comparison. As shown in the figure, the FE simulated results are about 30% lower compared to the measured values. This showed that the slotting process might have increased the iron loss.

![Graph showing iron loss measurements for non-slotted and slotted cores](image)

**Figure 3.14.** Measured and FE simulated AMM non-slotted and slotted cores iron loss without compensation.
3.3.2 Measurement Results in 50\(\text{Hz}\) to 1,000\(\text{Hz}\) Range

The AMM non-slotted core was machined to obtain a slotted machine stator. In order to quantify the change in iron loss caused by the cutting, the iron loss test was performed on the slotted stator core (see ”AMM3Meas” in Figure 3.15) and the iron loss density of the slotted and non-slotted cores were compared.

Nevertheless, the presence of the slots would disturb the uniform flux distribution of the core. The flux density along the yoke in circumferential direction is expected to be lowest under the stator teeth and highest under the stator slots due to the variations in axial length of the stator core. As iron loss is proportionally to the flux density squared, the regions of highest flux density are most important. In addition, iron loss is difficult to be compared due to the differences in weight between the non-slotted and slotted cores.

Based on the similar data (measured and calculated) as with the non-slotted core (assuming the flux is only concentrated within the yoke area hence neglecting any flux within the teeth), the yoke flux density (under the stator slots) and the iron loss density (dividing with the total core weight) found (“AMM3Meas”) is about two times lower than the non-slotted core (AMM1) as large areas of the yoke are operating at much lower value of flux density. Therefore, to obtain more accurate results the measured results were compensated by two first-order methods. The methods were based on using an effective ”average” measured flux density value (Compen1), and using 3D finite element model results to find an effective core weight (Compen2). The following described the compensation process.

1. Compen1: effective flux density, actual core weight

   • The measured core flux, found by integrating the sense coil voltage was divided with the area under the stator slots (yoke cross-sectional area) to obtain yoke flux density ”Meas \(B_y\)”.

   • Similarly, the flux was divided by the area under the stator tooth tip (tooth area under tooth tip + yoke cross-sectional area) to obtain the tooth flux density ”Meas \(B_t\)”.

   • ”Meas \(B_y\)” and ”Meas \(B_t\)” were averaged and the result used as the effective flux density.

   • The result is plotted as ”AMM3MeasAve” in Figure 3.15(a).
2. Compen2: peak flux density, effective core weight

- FE simulations were performed on the non-slotted core to obtain the iron loss density ($W/kg$) versus flux density ($T$) characteristic.
- Similar FE simulation were performed on the slotted stator with similar circumferential flux.
- Calculate the iron loss ($W$) as a function of peak flux density ($T$) under the stator slots ("FE By").
- At a similar peak flux density, estimate the effective stator core weight (44% of total weight of the AMM core) by dividing the loss density ($W/kg$) of non-slotted core with the iron loss ($W$) of slotted stator core.
- Convert the experimental iron loss measurements from the slotted stator core into iron loss densities with the effective weight from above and the peak flux density based on the area under the stator slots (yoke cross-sectional area).
- The result is plotted as ”AMM3MeasEffec” in Figure 3.15(b).

Figure 3.15 shows the iron loss density plot for the slotted stator core with the two compensation methods ("AMM3MeasAve", "AMM3MeasEffec") and without compensation ("AMM3Meas") compared to the non-slotted core ("AMM1Meas") results. The slotted stator core showed about two times lower iron loss without compensation compared to non-slotted core which is clearly unrealistic. With compensation method 1, a higher iron loss was obtained but it is still approximately 30% lower at 50Hz than the non-slotted AMM1 core (see Figure 3.15(a)). Nevertheless, at higher frequencies the iron loss density of both cores become comparable.

On the other hand, the slotted stator core exhibits 30% higher iron loss density compared to the non-slotted core with the compensation method 2 at high frequencies as shown in Figure 3.15(b). Under 200Hz, the losses were comparable.

Overall, compensation method 2 based on the effective weight obtained from finite element model is likely to provide the more accurate approach as it is based on FE analysis. Nevertheless, compensation method 1 gives reasonable accuracy if the finite-element model was not available.

The simulated iron loss densities for the non-slotted (JMAGRing) and the compensated slotted (JMAGStator) stator cores were plotted in Figure 3.16 for comparison.
Figure 3.15. Measured iron loss for AMM3 slotted stator core using various methods for compensating for the slotting, and AMM1 non-slotted core loss results at 50Hz to 1,000Hz range.
Figure 3.16. FE simulation results of non-slotted (JMAGRing) and slotted (JMAGStator) AMM stator cores using 2 different compensation methods.
The result based on compensation method 1 given in Figure 3.16(a) was consistent with the result shown in Figure 3.15(a). The slotted stator core showed smaller iron loss compared to the non-slotted core. On the other hand, the iron loss density of the non-slotted core was comparable with the slotted stator core compensated with the effective weight (Compen2) as shown in Figure 3.16(b). The result is as predicted, and hence verified the accuracy of the compensation method.

### 3.4 Core Loss Formula

The core loss of the stator can be estimated using the core loss values obtained from the non-slotted core iron loss tests conducted in Section 3.1. The simplest expression for calculating the iron loss is the well known Steinmetz equation (see Equation 3.7) which is the sum of the eddy-current loss and hysteresis loss for sinusoidal waveforms. The values for the unknown parameters \( k_e \) and \( k_h \) can be determined by curve fitting the available data from the iron loss curves. Then, the iron loss at any frequency and flux density can be estimated using the formula as given in Equation 3.7. Although, this iron loss formula is simple, it may not provide high accuracy estimation over a wide range of frequencies and flux densities.

\[
P_{\text{Fe loss}} = k_e \hat{B}^2 f^2 + k_h \hat{B}^2 f \quad (3.7)
\]

where \( k_e \): eddy-current loss coefficient (Steinmetz) and \( k_h \): hysteresis loss coefficient (Steinmetz).

A modified Steinmetz equation for calculating iron loss was introduced in [110] as given in Equation 3.8. For non-sinusoidal waveforms, the eddy-current term was modified to use rms value of the rate of change of flux density as shown in Equation 3.9.

\[
P_{\text{Fe loss}} = C_h f \hat{B}^{a+b\hat{B}} + C_e f^2 \hat{B}^2 \quad (3.8)
\]

\[
P_{\text{Fe loss}} = C_h f \hat{B}^{a+b\hat{B}} + C_e \left[ \frac{dB}{dt} \right]^2 \quad (3.9)
\]

where \( C_h \): hysteresis loss coefficient, \( C_e \): eddy-current loss coefficient, \( a, b \): constants and \( \frac{dB}{dt} \): rms value of the rate of change of flux density averaged over one cycle of the fundamental frequency.
Figure 3.17 shows the iron loss curves of the AMM ribbon at various magnetic flux densities and frequencies obtained by extrapolating the measured uncoated AMM1 non-slotted core iron loss curves shown previously in Section 3.2.2. The iron losses in an AMM stator can then be estimated from the iron loss curves for known tooth/yoke flux densities (see Section 5.2 and 5.3) and fundamental frequency (see Equation 5.41). A general expression for iron loss in terms of peak flux density ($\hat{B}$) and frequency ($f$) can be derived from the measured data extrapolated in the above curves.

![Iron Loss Curves](image)

**Figure 3.17.** Measured iron loss versus magnetic flux density curves of the uncoated AMM core (AMM1Meas).

In order to obtain the unknown coefficients, the core loss divided by frequency ($P/f$) graph for the minimum, mean and maximum operating flux densities for the designed machine were used. As for the frequency range, the iron loss coefficients should be optimised to give the best accuracy within the operating frequency range of the machine. The graphs were first curve fitted with straight lines, so that the intercepts and gradients can be used to obtain some of the coefficients and constants in Equation 3.8. Then, the remaining unknown parameters were obtained by solving the three simultaneous linear algebraic equations and taking average of the highest value for parameters with multiple solution values. Table 3.5 summarises values for each parameters for the prototypes constructed with uncoated AMM, grain-oriented SI and SMC.
Table 3.5. Iron loss formula coefficients for uncoated AMM, grain-oriented SI and SMC.

<table>
<thead>
<tr>
<th>Material</th>
<th>$k_e$</th>
<th>$k_h$</th>
<th>$C_h$</th>
<th>$C_e$</th>
<th>$a$</th>
<th>$b$</th>
</tr>
</thead>
<tbody>
<tr>
<td>AMM</td>
<td>$1.014 \times 10^{-5}$</td>
<td>$1.107 \times 10^{-2}$</td>
<td>$1.107 \times 10^{-2}$</td>
<td>$8.738 \times 10^{-6}$</td>
<td>2.883</td>
<td>0.714</td>
</tr>
<tr>
<td>SI</td>
<td>$2.345 \times 10^{-5}$</td>
<td>$6.605 \times 10^{-3}$</td>
<td>$6.605 \times 10^{-3}$</td>
<td>$2.184 \times 10^{-5}$</td>
<td>-0.0461</td>
<td>1.376</td>
</tr>
<tr>
<td>SMC</td>
<td>$1.432 \times 10^{-3}$</td>
<td>0.524</td>
<td>0.524</td>
<td>$1.662 \times 10^{-3}$</td>
<td>1.809</td>
<td>-0.308</td>
</tr>
</tbody>
</table>

Figure 3.18 gives the measured results and the curves calculated from Equation 3.8. As shown in the figure, the curves were matched closely. Also, the range of the flux densities chosen for curve fitting the plots can be varied according to the flux distribution in the stator core. Based on the coefficients extracted from the tests, the formula gives the specific loss (W/kg). The results can be multiplied with the respective weights of iron in the teeth and yoke to obtain the loss in Watts.

The iron loss approximation shown above did not take into consideration the flux in the radial direction within the core. The radial flux in an axial core was investigated in [121, 122]. The references provide models to evaluate the effect of radial flux and associated power loss in laminated cores. It was reported that the radial flux related loss is negligible compared to normal eddy-current losses if the core permeability (> 1000$\mu_0$), core conductivity (> $10^6$S/m) and number of poles (> 2) are high enough. Therefore, the radial flux related loss was not taken into consideration in this work.
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3.5 Intralaminar and Interlaminar Eddy-Current Losses

As it is known, the changing magnetic field induces opposing eddy-currents in the conductive iron core which produce resistive losses. The eddy-currents in laminated cores can be classified into two main types: intralaminar eddy-current loss due to eddy-currents within laminations (conventional eddy-current loss) and interlaminar eddy-current loss due to electrical contact between the laminations. The interlaminar eddy-currents can cause high iron losses as studied in [23], which is also observed in this research. Therefore, the modelling and analysis of the intralaminar and interlaminar eddy-current loss will be described in detail in the following sections.

3.5.1 Eddy-Current Loss Modelling

Firstly, three simplified models (EddyM1, EddyM2 and EddyM3) were derived to examine the eddy-current losses in the yoke of the machine assuming only circumferential flux exists, and that slotting is neglected so that the yoke has a uniform cross-sectional area. Then, the models were modified and combined to predict both the intralaminar and interlaminar loss. All the models assumed purely resistive current paths and a uniform flux density throughout the core which is unaffected by the presence of the eddy-currents.

The first eddy-current loss model (EddyM1) assumes a solid circular conductor of radius \( r_0 \) and length \( L \) to model the cross-sectional area of the core as shown in Figure 3.19(a). The eddy-currents can be assumed to flow in concentric circular paths within this cross-section.

The resistance \( R \) of a circular eddy-current path at radius \( r \) and width \( \Delta r \) (see Figure 3.19(a)) can be calculated from the electrical conductivity, \( \sigma \) of the material as given in Equation 3.10. Assuming a sinusoidally-varying magnetic flux density of peak value \( \hat{B} \) and angular frequency \( \omega \), the path is linked by the peak flux \( \hat{\Phi} \) given in Equation 3.11 and produces a peak back-EMF \( \hat{E} \) given by Equation 3.12. Hence, the average eddy-current loss per unit volume is estimated as a function of diameter \( D \) by integration as shown in Equation 3.13. The resulting eddy-current iron loss density expression shows the expected \( B^2 f^2 \) relationship and is also proportional to the material conductivity and the square of the diameter.
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\[ R(r) = \frac{1}{\sigma A} \frac{2\pi r}{\sigma L \Delta r} \]  
(3.10)

\[ \hat{\Phi}(r) = \pi r^2 \hat{B} \]  
(3.11)

\[ \hat{E}(r) = \omega \hat{\Phi}(r) \]  
(3.12)

\[ P_{EddyM1} = \left( \frac{4}{\pi D^2 L} \right) \int_0^{r_0} \frac{(\hat{E}(r))^2}{2R(r)} dr = \frac{\omega^2 \hat{B}^2 \sigma D^2}{64} \text{ W/m}^3 \]  
(3.13)

On the other hand, eddy-current loss model 2 (\textit{EddyM2}) assumes a solid square shape of side length \( l_x \) to model the cross-sectional area of the iron core as shown in Figure 3.19(b). A square eddy-current path through the material is assumed and the derived expressions are given in Equations 3.14-3.16 to calculate the loss per unit volume using a similar approach as in the model 1. Comparing the derived loss per unit volume in Equations 3.13 and 3.16 for the round and square cross-sectional areas, a ratio can be given as \( \frac{D^2}{l_x^2} \).
The model \( EddyM2 \) can be modified for anisotropic conductive materials by assuming a vertical conductivity of \( \sigma_1 \) and a horizontal conductivity of \( \sigma_2 \). Then, the loss can be calculated as given in Equations 3.17 to 3.19. Assuming that the interlaminar current path is in the horizontal direction and the intralaminar current path is in the vertical direction of model 2, and that the intralaminar conductivity is very much larger (\( \sigma_1 \gg \sigma_2 \)), Equation 3.19 can be simplified as in Equation 3.20. As a result, Equation 3.20 considers only the interlaminar eddy-current loss. To verify Equation 3.20, it can be seen that if only the interlaminar direction resistance (horizontal in Figure 3.19(b)) is accounted the resistance in Equation 3.14 would drop by a factor of two. This is equivalent to doubling the conductivity in Equation 3.16 as given in Equation 3.21, which is identical to Equation 3.20.

For eddy loss model 3 (\( EddyM3 \)), thin insulated laminations of width \( t_h \) within a square cross-sectional core are modelled (see Figure 3.20(a)). The resistance of the eddy-current path of a width \( \Delta x \) within the thin lamination at horizontal displacement \( x \) from the centre of the lamination is assumed to be only due to the vertical paths within

\[
R(x) = \frac{8x}{L\sigma\Delta x} \quad (3.14)
\]

\[
\tilde{E}(x) = \omega 4x^2\tilde{B} \quad (3.15)
\]

\[
P_{EddyM2} = \left( \frac{1}{P_i^2 \chi} \right) \int_0^{\frac{\Delta x}{2}} \frac{\tilde{E}(x)^2}{2R(x)} dx = \frac{\omega^2 \tilde{B}^2 \sigma_i l_x^2}{64} W/m^3 \quad (3.16)
\]
the lamination as shown in Figure 3.20(b). The expressions for the intralamination eddy-current loss are given in Equations 3.22-3.24 which is similar to \[123\]. Then, the total eddy-current loss is modelled as the combination of eddy loss \( P_{EddyM2b} \) with interlaminar conductivity \( \sigma_2 \) and \( P_{EddyM3} \) with intralaminate conductivity \( \sigma_1 \) as given in Equation 3.25.

(a) Diagram showing ten insulated laminations  (b) Eddy-current paths in a single lamination

**Figure 3.20.** Eddy-current loss model 3 for insulated laminated conductor.

\[
R(x) = \frac{2l_x}{L \sigma \Delta x} \quad (3.22)
\]

\[
\hat{E}(x) = \omega 2xl_x \hat{B} \quad (3.23)
\]

\[
P_{EddyM3} = \int_0^{t_h} \frac{1}{2R(x)} \left( \frac{\hat{E}(x))^2}{\hat{B}} \right) \frac{1}{l_xt_hL} \frac{1}{l_xt_hL} = \frac{\omega^2 \hat{B}^2 \sigma_1^2 t_h^2}{24} W/m^3 \quad (3.24)
\]

\[
P_{EddyM4} = \omega^2 \hat{B}^2 \left( \frac{\sigma_1 t_h^2}{24} + \frac{\sigma_2 l_x^2}{32} \right) W/m^3 \quad (3.25)
\]
3.5.2 Interlaminar Conductivity Measurement

The stator core of an axial-field machine consists of tape-wound AMM ribbon as shown in Figure 3.21. If the ribbon is perfectly insulated then the electrical resistance between the inner and outer diameter of the core is given by the total ribbon length, and its width, thickness and conductivity. This is defined as the intralamination resistance (see Figure 3.21(a)). The interlamination conductivity can be modelled in Figure 3.21(b) using an anisotropic conductive material which has infinite conductivity in the axial and circumferential planes, but low conductivity in the radial direction. This results in an electrical resistance (the interlamination resistance) which also exists between the inner and outer diameter of the core and is parallel to the intralamination resistance.

![Figure 3.21. Intralamination resistance and interlamination conductivity calculations.](image-url)
The interlaminar conductivity of the uncoated AMM case was estimated from the resistance of the toroidal core laminations. A four-wire resistance measurement was utilized to measure the resistance across the stack length ($SL$), (that is, between the inner and outer diameters) of the coated and uncoated non-slotted AMM toroidal cores as shown in Figure 3.21(a). The measured resistances are given in Table 3.6. It is deduced that the high resistance measured in the coated AMM core is the intralamination resistance and the small resistance in the uncoated AMM core is largely due to the interlamination resistance.

<table>
<thead>
<tr>
<th>Material</th>
<th>Measured $R$ (Ω)</th>
<th>Calculated Intralamination $R_{Lam}$ (Ω)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Uncoated AMM: Metglas 2605SA1</td>
<td>0.12</td>
<td>490</td>
</tr>
<tr>
<td>Coated AMM: Vitroperm 800</td>
<td>330</td>
<td>180</td>
</tr>
</tbody>
</table>

The intralamination resistance of the core is calculated using the fundamental resistance equation (see Equation 3.26). Figure 3.21(a) illustrates the parameters used in Equations 3.27 to 3.29. The calculated resistances using the material thickness and conductivity were included in Table 3.6. For the coated AMM core, the calculated value is about half the measured value. This is likely to be due to incorrect assumptions about the ribbon thickness and conductivity. Also, the much smaller measured resistance than the predicted intralamination resistance for the uncoated AMM showed consistency that the interlamination resistance was measured correctly.

$$R_{Lam} = \frac{l_{Lam}}{\sigma_1 A_{Lam}} \quad (3.26)$$

$$A_{Lam} = t_h x LW \quad (3.27)$$

$$Lam = \frac{SL}{t_h} (sf) \quad (3.28)$$

$$l_{Lam} = C_{Ave}(Lam) \quad (3.29)$$

where $l_{lam}$: total lamination length, $t_h$: thickness of lamination, $\sigma_1$: electrical conductivity of Metglas AMM ribbon (770,000 S/m), $sf$: stacking factor of lamination (see...
Equation 3.4), \( C_{Ave} \): average circumference of laminated toroidal core, \( A_{lam} \): laminated toroidal core cross-sectional area and \( Lam \): number of laminations.

Assuming that the measured resistance is purely due to interlamination conductivity \( \sigma_2 \), the latter can be estimated using Equations 3.30 and 3.31. The parameters in these equations are also defined in Figure 3.21(b). The value of \( \sigma_2 \) was calculated as 35.7 S/m and from this the interlaminar eddy-current loss can be calculated using the models given in Section 3.5.1.

\[
A_{LW} = \frac{2\pi R_{StaOut} + 2\pi R_{StaIn}}{2} \quad (3.30)
\]

\[
\sigma_2 = \frac{SL}{RA_{LW}} \quad (3.31)
\]

### 3.5.3 Result Comparisons

Equation 3.32 gives the ratio between the equation for power loss (\( P_{EddyM4} \)) due to both intralamination and interlamination conductivities against the classical thin lamination power loss equation (\( P_{EddyM3} \)). Note that the power loss is independent of frequency and flux density, and is proportional to the ratio of the interlamination to intralamination conductivities and the square of the ratio of the cross-sectional core width to the lamination thickness. For the uncoated AMM core with the estimated interlaminar conductivity from the previous section, the ratio is about 50 using the values shown in Table 3.7.

\[
\frac{P_{EddyM4}}{P_{EddyM3}} = 1 + \frac{3}{4} \frac{\sigma_2}{\sigma_1} \frac{l_x^2}{h^2} \quad (3.32)
\]

Table 3.8 shows the comparison between the prediction of the classical eddy-current power loss equation (\( P_{EddyM3} \)) for thin laminations and the data sheet values. The table is given for the total iron loss of uncoated AMM 2605SA1 at various frequencies where available. The ratio is expected to decrease with increasing frequency as eddy-current becomes more dominant with regards to hysteresis loss. As it can be seen in the table, the ratio is the highest (about 200) at 50Hz which is reduced to 5 (about 40
Chapter 3 AMM Iron Loss Measurements

Table 3.7. Parameter values of the model.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Intralamination conductivity $\sigma_1$</td>
<td>770,000 S/m</td>
</tr>
<tr>
<td>Interlamination conductivity $\sigma_2$</td>
<td>35.7 S/m</td>
</tr>
<tr>
<td>Lamination thickness $t_h$</td>
<td>0.0254 mm</td>
</tr>
<tr>
<td>Core width $l_x$</td>
<td>30 mm</td>
</tr>
<tr>
<td>$\frac{P_{\text{EddyM4}}}{P_{\text{EddyM3}}}$</td>
<td>50</td>
</tr>
</tbody>
</table>

times less) at 1,000Hz. The high ratio indicates that classical eddy-current loss forms a low fraction of the total iron loss for this very thin (0.0254 mm) AMM material.

In addition, the measured iron loss is normalised with the total eddy-current loss prediction ($P_{\text{EddyM4}}$) for various frequencies and given in the table. The ratio is highest at 50Hz at about 40 and reduces to about four times at 1,000Hz. This result indicates that the calculated interlamination resistance eddy-current losses is about 25% of the measured AMM core losses at high frequencies.

Table 3.8. Comparison of data sheet, measured iron loss and calculated loss of uncoated AMM.

<table>
<thead>
<tr>
<th>Frequency (Hz)</th>
<th>Datasheet DS (mW/kg)</th>
<th>Measured Meas (mW/kg)</th>
<th>Model 3 $P_{\text{EddyM3}}$ (mW/kg)</th>
<th>Model 4 $P_{\text{EddyM4}}$ (mW/kg)</th>
<th>$\frac{P_{\text{EddyM4}}}{P_{\text{EddyM3}}}$ ratio</th>
<th>$\frac{P_{\text{DS}}}{P_{\text{EddyM4}}}$ ratio</th>
</tr>
</thead>
<tbody>
<tr>
<td>50</td>
<td>64</td>
<td>550</td>
<td>0.3</td>
<td>14</td>
<td>213</td>
<td>39</td>
</tr>
<tr>
<td>60</td>
<td>80</td>
<td>-</td>
<td>0.4</td>
<td>20</td>
<td>200</td>
<td>-</td>
</tr>
<tr>
<td>100</td>
<td>-</td>
<td>1,250</td>
<td>1.1</td>
<td>56</td>
<td>-</td>
<td>22</td>
</tr>
<tr>
<td>200</td>
<td>-</td>
<td>2,800</td>
<td>4.5</td>
<td>230</td>
<td>-</td>
<td>12</td>
</tr>
<tr>
<td>400</td>
<td>1,500</td>
<td>-</td>
<td>18</td>
<td>900</td>
<td>83</td>
<td>-</td>
</tr>
<tr>
<td>500</td>
<td>-</td>
<td>7,500</td>
<td>28</td>
<td>1,400</td>
<td>-</td>
<td>5</td>
</tr>
<tr>
<td>1,000</td>
<td>5,000</td>
<td>21,700</td>
<td>110</td>
<td>5,600</td>
<td>45</td>
<td>4</td>
</tr>
</tbody>
</table>
3.6 Conclusions

In order to quantify the magnitude of iron losses at various frequencies for the selected magnetic materials, iron loss measurements were conducted. The test results showed that at $50\,\text{Hz}$, coated AMM has the lowest iron loss followed by grain-oriented SI, uncoated AMM, non-oriented SI and then SMC.

In this research, uncoated AMM and grain-oriented SI were utilised to construct the prototype axial-flux machines. It was found that at $50\,\text{Hz}$ the uncoated AMM has 50% higher measured losses than the grain-oriented SI. However the losses are comparable at $200\,\text{Hz}$, and uncoated AMM has 30% lower loss at $1,000\,\text{Hz}$.

It was suspected that the factor of four times increased loss measured in the uncoated AMM versus its datasheet values at high frequencies is due to interlaminar eddy-current loss. Three models were derived to investigate the intralaminar and interlaminar eddy-current loss. It was found that the calculated ratio of interlaminar to intralaminar eddy-current loss is proportional to the interlaminar conductivity and the core dimensions. This ratio was about 50 for the core tested. At high frequencies, the calculated interlaminar eddy-current loss accounted for only about 25% of the measured iron losses indicating that further work is required to explain the large discrepancy between the measured and datasheet iron losses.

Two compensation methods were implemented to estimate the iron loss density of slotted stator cores based on the results of the circumferential flux test. The first method was based on the average flux density in the slotted core and the total core weight. The second method used the peak yoke flux density in the core and a reduced effective core weight that was determined by FE analysis to match the iron loss densities between the non-slotted and slotted cores. It was found that the slotted stator core has 30% more iron loss density than the non-slotted stator core which is likely to be caused during the cutting process.

The measured iron loss data for uncoated AMM was utilised in the 3D finite element modelling. In this study, the measured and simulated results for the non-slotted core matched closely, which gives confidence in the modelling approach.

It should be noted that the results presented in this chapter are also utilised in Chapter 6 to model the AFPM machine to examine its iron loss characteristics. In addition, based on the measured iron loss data, the iron loss coefficients were extracted for analytical analysis of the machine in Chapter 5.
Chapter 4

32mm Machine Analytical Design and Experimental Results

This chapter describes a small size (32mm outer diameter) concept demonstration motor design to examine the benefits of an AFPM structure based on cut AMM and SMC materials. The machines were constructed and the electrical characteristics of the motor configurations were investigated. 3D FEA was utilised to model and analyse the AFPM motors. In addition, experimental tests were conducted to investigate their performance characteristics. These include the effects of back-iron, magnet shape and airgap length on machine performance. A direct comparison of the AMM machine with SMC is also studied in this chapter.
4.1 Machine Design and Analysis

A three-tooth concept demonstration machine was used in the initial phase of the research to investigate the feasibility and characteristics of axial-field AMM machines. It should be noted that the stator core was designed, cut and assembled by the industrial partner for a selected application that is commercially sensitive. The stator core was not optimised for this prototype machine and demonstrates a poor utilisation of active material. In addition, it was cut from an uncoated AMM toroid. Therefore, as explained in Chapter 3 higher iron loss compared to manufacturer’s data is expected. Figure 4.1 gives the dimensions of this cut stator core.

![Dimensions of the stator core.](image)

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Outer Diameter</td>
<td>32mm</td>
</tr>
<tr>
<td>Inner Diameter</td>
<td>11mm</td>
</tr>
<tr>
<td>Slot Depth</td>
<td>8 mm</td>
</tr>
<tr>
<td>Yoke Thickness</td>
<td>5 mm</td>
</tr>
<tr>
<td>Slot Width Angle</td>
<td>70°</td>
</tr>
</tbody>
</table>

Figure 4.1. Dimensions of the stator core.

The stator core of the AMM machine (see Figure 4.2(a)) was cut using an abrasive water jet cutting technique [26] and for comparison purpose an identical SMC stator core (see Figure 4.2(b)) was fabricated by machining.

In these prototype designs, a concentrated wye-connected three-phase windings were used. As shown in Figure 4.2(c), the three-phase coils were wound around the three stator teeth, and a two pole rare-earth magnet rotor was constructed. Based on the specifications of the stator, inverter and dynamometer setup, the analytical design was performed using the following specifications:

- Speed: 6,000 rpm
- Line to line voltage (peak): 25 V
- Number of poles (2p): 2
- Copper packing factor: 30% (typical 20-60%)
- Copper current density $J = 3A_{rms}/mm^2$ (typical 2 – 5A$_{rms}$/mm$^2$)
4.1.1 Analytical Design Approach

In order to obtain the number of turns in each phase winding that is located around each tooth, the induced voltage equation is used, which is given by

\[ E = 4.44 N_{ph} \Phi_{pk} f \]  \hspace{1cm} (4.1)

where \( E \): rms induced voltage, \( f \): supply frequency, \( \Phi_{pk} \): peak flux in the stator teeth and \( N_{ph} \): number of turns per phase.

In Equation 4.1, the value of induced voltage and the frequency were obtained from the specifications chosen. The airgap flux density was calculated based on the magnet remanent flux density \( B_r \), the magnet thickness \( l_m \) and airgap length \( l_g \) as given in Equation 4.2.

\[ B_g = \frac{l_m}{l_m + l_g} B_r \]  \hspace{1cm} (4.2)

In this equation, it is assumed that the airgap flux density is uniform over the stator teeth and is zero over the slots. From the calculated flux density, the peak flux can be estimated using the stator tooth area, \( P \) (see Figure 4.4).

The magnet length in this study was based on the available commercial magnets (Grade N38 with 1.3T max \( B_r [124] \)). The airgap length was chosen based on the mechanical requirements (due to the high axial force between the magnets and stator) and the available slot area to accommodate the required number of turns to satisfy the speed and voltage requirements in Equation 4.1.

**Figure 4.2.** The stator cores of the AFPM test motors a) using cut AMM ribbon, b) using SMC c) AMM stator with windings.
Figures 4.3(c) and 4.3(b) show the simplified cross-sectional views of two possible AFPM machine configurations with and without back-iron respectively. The function of the back-iron in Figure 4.3(c) is to improve the airgap flux density. Note that the back-iron is stationary as it models the effect of a double-sided machine configuration (two stators and one rotor). The designs in these figures are used in the analytical study of the sample AMM cut stator based AFPM machines. For the configuration without back-iron, the total airgap length was roughly approximated as twice the length of the configuration with back-iron in the calculation.

![Simplified cross-sectional views of three-phase AFPM motor configurations](image)

**Figure 4.3.** Simplified cross-sectional views of three-phase AFPM motor configurations a) with back-iron, b) without back-iron, c) sketch of the PM rotor.

In the practical machine, it is required to have a reasonable size rotor shaft (diameter) for the design speed range. However as it can be seen in Figure 4.2, due to lack of room available inside the inner diameter of the stator, the cross-sectional area of the coils has to be limited to still maintain the required shaft size. As expected, this severely limits the achievable electric loading and hence the output torque in such machine configuration.

For the winding design, the diameter of the wire was determined based on the number of turns calculated using Equation 4.1 and the available slot winding area as highlighted in Figure 4.4. As shown, the slot area was estimated assuming a rectangular coil of half of width 'W' and of height 'H'. Then, the total copper area was estimated using the packing factor chosen (30%). In the analytical design, the required number of turns was used to determine the wire diameter.
The phase resistance of the stator windings was calculated based on Equation 4.3 by estimating the total length and cross-sectional area of the wire. A simple reluctance $R$ and inductance $L$ estimations were implemented at this stage as given in Equations 4.4 and 4.5.

$$R_{wire} = \frac{l_{wire}}{\sigma_{cu} A_{wire}} \quad (4.3)$$

$$R = \frac{l_g + l_m}{A_{tooth} \mu_o} \quad (4.4)$$

$$L = \frac{N_{Ph}^2}{\frac{R}{\mu}} \quad (4.5)$$

where $l_{wire}$: total length of wire per phase, $A_{wire}$: wire cross-sectional area, $\sigma_{cu}$: electrical conductivity of copper \(59 \times 10^6 \text{Sm}^{-1}\), $\mu_o$: permeability of free space and $A_{tooth}$: tooth cross-sectional area.

The iron loss per unit weight ($\text{W/kg}$) was approximated based on Equation 3.7 with the coefficients derived from the measured data as discussed in Chapter 3 and the estimated peak flux densities were obtained using Equation 4.2 (see Table 4.1). The final iron loss value in watts can be obtained using the weight of the stator.

To find the output power of the machine, the phase induced voltage and current were calculated first. The phase voltage was estimated using Equation 4.1 and the maximum allowable phase current ($I_{PhMax}$) was determined from the wire’s current density as given in Equation 4.6. Then, the shaft torque was estimated from the output power and speed.
\[ I_{PhMax} = JA_{wire} \]  

where \( J \): current density assume \( 3A_{rms}/mm^2 \).

A summary of the results of the analytical design for the AMM stator (Figure 4.2(a)) are given in Table 4.1. It was estimated that the two-pole machine with back-iron “2 pole(BI)” at 6,000 rpm produces 4.4 mNm of output torque to give an output power of 2.74 W at a maximum phase current of about 0.1 A per phase and an efficiency of about 92%. For the same setup without back-iron “2 pole”, it was calculated that the torque will drop to 3.1 mNm, which corresponds to an output power of 1.96 W, at a maximum phase current of about 0.1 A and an efficiency of about 90%.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>2 pole</th>
<th>2 pole(BI)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnet remanent flux density, ( B_r(T) )</td>
<td>1.30</td>
<td>1.30</td>
</tr>
<tr>
<td>Phase terminal voltage ( V(V) )</td>
<td>6.94</td>
<td>9.72</td>
</tr>
<tr>
<td>Length of total airgap ( l_g(mm) )</td>
<td>4</td>
<td>2</td>
</tr>
<tr>
<td>Length of magnet ( l_m(mm) )</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>Peak flux ( \Phi_{pk}(mWb) )</td>
<td>0.055</td>
<td>0.077</td>
</tr>
<tr>
<td>Rated frequency ( f(Hz) )</td>
<td>100</td>
<td>100</td>
</tr>
<tr>
<td>Airgap flux density, ( B_g(T) )</td>
<td>0.56</td>
<td>0.78</td>
</tr>
<tr>
<td>Number of turns per phase, ( N )</td>
<td>285</td>
<td>285</td>
</tr>
<tr>
<td>Radius of wire ( r_{wire}(mm) )</td>
<td>0.100</td>
<td>0.100</td>
</tr>
<tr>
<td>Resistance of wire ( R_{wire}(\Omega) )</td>
<td>8.2</td>
<td>8.2</td>
</tr>
<tr>
<td>Inductance, ( L(mH) )</td>
<td>0.77</td>
<td>0.82</td>
</tr>
<tr>
<td>Max current per phase, ( I_{PhMax}(A) )</td>
<td>0.094</td>
<td>0.094</td>
</tr>
<tr>
<td>Output power, ( P_M(W) )</td>
<td>1.96</td>
<td>2.74</td>
</tr>
<tr>
<td>Torque, ( T_M(mNm) )</td>
<td>3.1</td>
<td>4.4</td>
</tr>
<tr>
<td>Copper loss, ( P_{copper loss} = 3I_{PhMax}^2R_{wire}(W) )</td>
<td>0.217</td>
<td>0.217</td>
</tr>
<tr>
<td>Iron loss, ( P_{iron loss}(W) )</td>
<td>0.010</td>
<td>0.025</td>
</tr>
<tr>
<td>Efficiency, ( \eta(%) )</td>
<td>89.6</td>
<td>91.9</td>
</tr>
</tbody>
</table>
Tooth Magnetic Flux Waveform

According to [110], the back-waveform can be estimated analytically by calculating the airgap flux density distribution and hence the stator tooth flux as a function of rotor position. The back-EMF is then obtained by differentiating the tooth flux-linkage (product of tooth flux and the number of turns) waveform with time.

To determine the airgap flux distribution in the machine for a given rotor position a simple analytical method was implemented. Figure 4.5 shows the “unrolled” machine structure illustrating the stator teeth and rotor magnets. An idealised representation of the airgap flux density waveform can be calculated for a given rotor position, taking into account that the flux entering the rotor must equal the flux leaving it. The flux in each tooth can be obtained by integrating the flux density across the tooth.

![Diagram illustrating the "sliding" concept by comparison of the location of stator teeth and rotor magnets.](image)

The calculation was performed numerically by representing the stator teeth and rotor magnets as arrays which could be multiplied together to obtain the airgap flux density waveform. This was then adjusted to meet the zero net flux requirement described above. The rotation of the rotor magnet was simulated by virtually sliding the rotor magnet array relative to the stator tooth array. In this way, the tooth flux-linkage as a function of rotor angle can be obtained. Finally, the tooth induced voltage was obtained by taking the rate of change of tooth flux-linkage with time.
The rotor of the test machines was constructed with neodymium disc magnets of 12\textit{mm} diameter (Rotor 1, in Figure 4.9(a)). In the analytical model, the disc magnets were simplified as arc-shaped magnets of similar surface area as shown in Figure 4.6.

**Figure 4.6.** Analytical tooth magnetic flux waveform modelling based on arc-shaped PMs on the rotor.

The analytical, FE and measured phase back-EMF plots are shown in Figure 4.7. In these waveforms, the dip that occurs near the waveform peak is due to the 180° spacing between the two rotor magnets (see Figure 4.9). The plots showed similar shapes but rounder edges were observed for the measured and FE simulated waveforms. The rather crude analytical back-EMF waveform is due to the magnet geometry simplification and the lack of fringing in the modelling method. However, the analytical method still provides a rough approximation of the magnitude and the general shape of the back-EMF waveform. The analytical method could readily be improved in accuracy if desired by estimating the fringing.

**Figure 4.7.** Analytical (Anal), FE simulated (FE) and measured (Meas) phase back-EMF plots.
4.1.2 3D FEA Modeling

The finite element method is a numerical method which divides the original field problem domain into smaller sub-domains called finite elements. The potential distribution in each element is represented by partial differential equations derived from Maxwell’s equations. The potential distribution in the problem can be determined by solving the respective equations using a computer.

Finite element software can be used to predict the designed machine’s behavior more accurately than analytical models as it avoids the need for many of the simplifying assumptions with regards to flux distributions, geometric arrangement and lack of fringing and saturation. Hence, it is often used in verifying the analytical results and the design concept of the machine before the construction stage. The details of the simulation for the parameters and performance under time-stepping inverter operation will be described in Sections 6.3.1 and 6.3.2.

Although a machine’s symmetry can be used to reduce the FE model size, a complete machine model of the cut AMM-based test motor was implemented here due to the lack of symmetry in the test machine. In addition, because of the 3D geometry of the axial-field PM machine and the limited capabilities in the FEA package (JMag) 3D geometry editor, the 3D model was first generated using the Solid Edge CAD package and later imported into JMag. Figure 4.8 illustrates a sample 3D mesh (with 13,937 nodes and 82,572 elements) and views of the magnetic flux density vectors and contour plot of the cut AMM stator with Rotor 1 (disc magnets of 12mm diameter).

(a) 3D Mesh view    (b) B vector display, (T)    (c) B contour plot, (T)

Figure 4.8. 3D FE model of the AMM axial field test motor (without back-iron). Note that the non-magnetic holder is not shown in 4.8(b) and 4.8(c).
Chapter 4 Machine Analytical Design and Experimental Results

4.2 Construction and Testing

4.2.1 Construction of the Motor

The results of the analytical design given in Table 4.1 were utilised to construct the two-pole AMM-based AFPM test motor. As shown in Figure 4.9(a) that the AMM stator contains three coils wound (one for each phase) around the three teeth, a rotor with two pole magnets and the AMM back-iron. Based on the available commercial magnets, two other rotors were built as shown in Figure 4.9(b).

![Test motor components](image)
![Rotor magnets configurations](image)

Figure 4.9. Photographs of the three-phase AMM stator with windings, AMM back-iron and three different 2 pole rotor designs: Rotor 1 (with 12mm disc shape magnets), Rotor 2 (with 20x12.5mm block magnets) and Rotor 3 (with 15mm disc shape magnets).

In this study, a tape-wound 4mm thick uncoated AMM disk was initially used as the back-iron. However, the AMM disk could not withstand the large axial magnetic forces during the test. Therefore, it was replaced by a solid piece of mild steel. Although this approach improved the magnitude of the back EMF, it would increase the eddy-current losses significantly compared to the design without back-iron. This was investigated in Section 4.3.5.
4.2.2 Custom-built Test Rig

A setup was designed for a wide range of interchangeable axial-field stator and rotor configurations (e.g. single, double stator and alternative back-iron configurations) and to allow easy variation of the airgap length. Figure 4.10 shows the custom-built test setup used in this research. The setup includes a high-speed DC machine (ASTRO - 15 Super Ferrite, 15,000rpm, 12Vdc) for loading the AFPM motor configuration under test. A special attachment introduced in the test setup allows the variation of the axial airgap length from 0.5 mm up to 10 mm with an accuracy of 0.1 mm. In addition, the setup was designed to achieve good stability and rigidness during high speed operation and can accommodate axial-field motors up to 80 mm in diameter.

![Test rig used with the AFPM motor with adjustable airgap and the dc load machine. The airgap has been set to a large value for display purposes.](image)
4.2.3 Motor Drive

A three-phase motor drive hardware was also implemented to test the designed AFPM machine as a motor, which has the principal block diagram shown in Figure 4.11. The rotor position in this drive was measured by three Hall-effect sensors, which were excited by a small custom-built PM disk.

![Figure 4.11. The AFPM motor drive hardware implementation. A photograph of the motor drive system is given in Appendix A.](image)

The Microchip PIC16F877A microcontroller was used to implement the motor controller. This CMOS FLASH based 8-bit microcontroller is capable of 200\(\text{ns}\) instruction execution and is programmed using only 35 simple single-word instructions.

Based on the rotor position, the appropriate stator winding phases were energised. In a star-connected three-phase winding, two phases are energised at any one time. For the microcontroller implementation, a state table with the list of possible output drive codes was derived, and the sensor outputs were used as the table offset pointer [125].

A voltage source inverter with a pulse-width modulation (PWM) technique was used to control the winding current, which varied motor speed. The duty cycle of the PWM was adjusted using a reference voltage that was sampled via an analogue-to-digital converter (ADC).

The microcontroller generated six PWM control signals (one for each of the inverter switches), which were isolated via opto-couplers. An integrated power module, International Rectifier IRAMX16UP60A (16A, 600V) was used in the inverter which contains free-wheeling diodes, gate drivers, level shifters as well as an over-temperature/over-current protection circuit.
Drive System Testing

Preliminary testing of the power module and control system was performed using simulated Hall-effect sensor output signals that were connected to the controller. This was achieved via an oscillator that generates three square-wave signals with a $60^\circ$ phase shift between them.

To test the timing accuracy and synchronisation of the position sensors, firstly the inverter output was connected to a star-connected resistive load. These simulated sensor signals and a set of output phase voltage waveforms across the resistive load are shown in Figure 4.12. The measured results were used to check the commutation timing between the sensor outputs and the generated motor voltages.

![Figure 4.12. Resistive load test: measured A-phase voltage (top) and three simulated Hall-effect sensor signals. Vertical scale 5 V/div, horizontal scale 0.01 s/div.](image)

After the timing and synchronisation tests, the inverter and the control system were connected to the AFPM test machine. Figure 4.13 shows a set of typical measured phase voltage and current waveforms at 100% PWM duty cycle which resulted a maximum speed of 5,700 $rpm$ in the motor (close to the design speed of 6,000 $rpm$) at no-load condition. The distorted current waveform under no-load conditions is due to the combination of the non-sinusoidal back-EMF waveform (see Figure 4.13) and the use of a square-wave voltage-source inverter.
It was observed that the maximum achievable no-load speed using the drive system varies with the type of dc generator attached on the shaft. A maximum speed of 5,700 rpm was obtained with an ESCAP dc machine (low inertia ESCAP34L11, 5,390 rpm, 24 Vdc) and 3,500 rpm with an ASTRO-15 (15,000 rpm, 12 Vdc) dc machine. The test motor’s input current was significantly higher with the ASTRO-15 due to lower back-EMF constant. It was also observed that when ESCAP machine was used to drive the rotor shaft of the test motor from standstill, a significant high torque is required to overcome the high cogging torque of the test machine. This effect was especially important for small airgap lengths and for the AMM-based stator which has a higher permeability than the SMC based stator. Therefore, the ASTRO-15 motor was used to conduct the generator tests of the test machine.

4.2.4 Experimental Procedures

The dc resistance was measured by the ammeter/voltmeter method using a dc power supply. The inductance measurement was carried out across two of the motor terminals while two of the phases connected in parallel as shown in Figure 4.14. This connection arrangement takes into account the mutual coupling between all three phases, and gives an inductance equal to 1.5 times the phase inductance.
The back-EMF constant of the test machine was obtained by plotting the induced phase voltage against the machine speeds and obtaining the slope of the curve that is equal to the back-EMF constant of the test motor.

![Figure 4.14. Three phase winding connection for inductance measurement.]

In addition, the open-circuit loss was also obtained using the dc motor to drive the test machine. The dc motor was first tested at no-load. Then, the dc motor was used to drive the brushless AFPM machine under open-circuit condition. The difference gives the open-circuit losses of the brushless AFPM motor (see Figure 4.15).

![Figure 4.15. Measured open-circuit losses in the test set up.]

For inverter testing, the load on the test machine was varied by changing the resistance on the dc loading generator. The output torque was then calculated utilising the dc generator as the torque transducer (see Section 7.3.2), which also considered the losses measured due to the absence of a torque transducer. It should be noted that all the measured torque and efficiency shown in this thesis were estimated using the method as will be described in details in Section 7.3.2.

The detailed procedures and calculations will be described in Section 7.3.
4.3 Results - Parameters of the Test Motor

4.3.1 Resistance and Inductance

The analytical, FE and measured results are summarised in Table 4.2. The results demonstrate that the winding resistance matched closely, but the measured and FE based inductance is about an order of magnitude larger than the analytical value. This is due to the fact that the simple analytical calculation utilised is not a good approximation due to the large airgap in the design and the high leakage flux associated with the three-dimensional (3D) salient-pole stator structure. The FE prediction closely matches the measured inductance value.

Table 4.2. Analytical, FE simulated and measured AFPM test motor parameters, Rotor 1 (12mm disc shape magnets)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Analytical</th>
<th>FE</th>
<th>Measured</th>
</tr>
</thead>
<tbody>
<tr>
<td>Winding resistance (Ω)</td>
<td>8.2</td>
<td>9.0</td>
<td>8.9</td>
</tr>
<tr>
<td>Winding inductance (mH)</td>
<td>0.77</td>
<td>5.3</td>
<td>5.1</td>
</tr>
</tbody>
</table>

4.3.2 Back-EMF Profile

In [126], it was shown that the harmonic content of the induced voltage is strongly affected by the rotor magnet design even though different PM rotor shapes showed similar overall performance in an AFPM generator. In this section, the back-EMF profiles of rotors with different magnet shapes and sizes as shown in Figure 4.9 are investigated. The performance of the test motor with these rotor configurations are shown later in Section 4.4.1.

Figure 4.16(a) shows the back-EMF waveforms of the three rotor configurations (Rotor 1, 2 and 3). The results were obtained under identical operating conditions: at a speed of approximately 4,000 rpm and an airgap of 1mm. In the back-EMF waveforms, the dip observed near the waveform peak is due to the large spacing between the two rotor magnets. The test results revealed that the magnitude of the back-EMF waveform increases as the magnet surface area increases.
Figure 4.16(b) shows a comparison of the FE simulated and measured back-EMF waveforms for 1mm and 4mm airgap lengths for Rotor 1. As can be seen, there is a good correspondence between the simulated and measured results which gives confidence in the accuracy of the 3D finite element modeling approach used. The small discrepancies in their amplitudes are likely to be due to the asymmetry in the rotor magnet positions.

![Image](a)

![Image](b)

Figure 4.16. a) Measured phase back-EMF waveforms for three rotor configurations of Rotor 1, 2 and 3, b) FE simulated and measured phase back-EMFs at 4,000rpm and 1mm airgap with Rotor 1.

The FE simulated tooth flux and back-EMF waveforms with and without back-iron are given in Figure 4.17. The waveforms were obtained with zero stator excitation in the 3D FE tool. As predicted, higher peak flux density was obtained with the presence of the back-iron and in this case it was about a 30% increase. As shown in Figure 4.17(b), the motor configuration with back-iron generates a higher back-EMF voltage due to the increased airgap flux density.

The analytical, FE and measured phase back-EMF magnitudes with and without back-iron are given in Figure 4.18(a). It was observed that the measured back-EMF voltage with back-iron was about 10% and 7% lower than the analytical and FE values which may be due to the asymmetry of the rotor magnets resulting to different peak amplitude. In addition, as expected the magnitude of phase back-EMF increases linearly with speed and decreases as the airgap length increases (see Figure 4.18(b)).
4.3.3 Axial Force and Cogging Torque

Figures 4.19 and 4.20 provide the FE simulated plots for the axial force and cogging torque. Due to hardware availability (e.g. torquemeter), experimental confirmation of these results were not obtained.

Figure 4.19(a) shows that the single-sided configuration with no back-iron has large axial forces (about 15N for a 1mm airgap). This could produce large bearing losses.
In order to reduce the axial force, a symmetrical double-sided stator design (i.e. two stators, one on each side of the disc rotor) can be utilised. Adding back-iron reduces the force by about a quarter but also changes the direction of the net force. In addition, the average axial force decreases as the airgap length increases due to higher flux leakage as shown in Figure 4.19(b).

![Instantaneous Axial Force Graph](image1)

(a) FE results of instantaneous axial force

![Average Axial Force Graph](image2)

(b) FE result of average axial force as a function of airgap lengths

**Figure 4.19.** a) FE simulated instantaneous axial force as a function of rotor angle at 1\text{mm} airgap with Rotor 1, b) average axial force as a function of airgap length with Rotor 1.

![Cogging Torque Graph](image3)

**Figure 4.20.** FE simulated cogging torque as a function of rotor angle at 1\text{mm} airgap with Rotor 1.

The predicted cogging torque as a function of rotor position is given in Figure 4.20. The interaction between each of the three stator teeth and two magnet poles gives six pulses per revolution. The peak cogging torque is about 3 times of the rated machine torque (4.4\text{mNm}). This large value is perhaps not surprising given the machine only has three...
teeth. In addition, the waveforms show that the cogging torque was not significantly affected by the presence of the back-iron.

### 4.3.4 Iron and Open-Circuit Losses

Table 4.3 gives the analytical and FE peak tooth flux density and stator iron loss predictions. The FE model predicts about 6% higher tooth flux density which could be due to fringing flux entering the sides of the teeth. It predicts to about 60-70% higher iron loss. In addition, the presence of back-iron increases the tooth flux density about 1.4 times. As the iron loss is proportional to flux density squared, twice the iron loss was obtained with the presence of back-iron.

<table>
<thead>
<tr>
<th>Method</th>
<th>Peak Tooth Flux Density (T)</th>
<th>Iron Loss (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>W/O back-iron</td>
<td>W/ back-iron</td>
</tr>
<tr>
<td>Analytical</td>
<td>0.56</td>
<td>0.78</td>
</tr>
<tr>
<td>FE</td>
<td>0.60</td>
<td>0.83</td>
</tr>
</tbody>
</table>

Figure 4.21(a) gives the FE simulated iron loss with and without back-iron for no-load and loaded conditions. Between 7 to 15% increase in iron loss was predicted when the test motor is loaded. This is due to the increase in the stator flux density due to the three-phase currents.

Figure 4.21(b) gives the measured open-circuit loss as a function of speed with and without back-iron. The measured open-circuit losses (including windage, bearing and iron losses) are considerable larger than the FE simulated iron loss. From the tests, the open-circuit loss with back-iron at 3,000rpm is just under 3.2W. The FE simulated iron loss at 3,000rpm was 0.016W, twenty times smaller. In addition, the calculated output power of the two-pole machine is only about 1.4W (half the speed and values in Table 4.1). This large loss was not unexpected given the use of a solid back-iron. However, it was not expected to have open-circuit loss of 0.67W, which is about 85 times higher than the FE simulated iron loss.

A possible explanation for the high losses is the large axial forces associated with the axial field topology causing large bearing losses. The 3D FEA indicates a significant axial force of about 15N with the 1mm airgap for the case without rotor back-iron (see Figure 4.19).
Another possible explanation is the presence of magnet and back-iron eddy-current losses, which will be studied in the next section, where 3D FEA was utilised to simulate such losses. Further investigation on axial bearing and other open-circuit losses were conducted in Chapter 8 using a larger size AFPM machine which uses an improved test setup with selected bearings to reduce the bearing loss.

![Iron loss, no-load (NL) and load (L) case](image1)

![Measured open-circuit loss](image2)

**Figure 4.21.** a) Analytical and FE simulated iron loss for no-load (NL) and loaded (L) conditions with Rotor 1, b) measured open-circuit loss with and without back-iron with Rotor 1.

The magnitudes of the various open-circuit loss components (measured or calculated) are given in Table 4.4. As described in Section 4.2.4, a dc motor was utilised to measure the open-circuit loss (including bearing and windage losses) based on measuring its input current and using the knowledge of the dc motor’s torque constant and open-circuit loss (see Figure 4.15). The accuracy of the measurement is mainly affected by variations in the dc motor open-circuit loss. An error analysis indicates an accuracy of ±0.02W for the loss measurement at 3,000rpm.

The main components of such losses are the measured bearing loss due to radial loading and the solid back-iron eddy-current loss. By subtracting the calculated or measured loss components from the total measured loss, the unaccounted loss was 0.4W for without back-iron and 1.8W for with back-iron configuration. Although, a significant portion of this unaccounted loss could be due to axial loading of the bearings, however it is not clear why such losses are so different while the axial loadings are comparable.
Table 4.4. Open-circuit loss components of the small AFPM test motor at 3,000 rpm with Rotor 1

<table>
<thead>
<tr>
<th>Loss component</th>
<th>W/O back-iron (W)</th>
<th>W back-iron (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Total measured open-circuit</td>
<td>0.59</td>
<td>3.20</td>
</tr>
<tr>
<td>Radial bearing loading (measured)</td>
<td>0.13 ± 0.02</td>
<td>0.13 ± 0.02</td>
</tr>
<tr>
<td>Windage (measured)</td>
<td>0.04 ± 0.02</td>
<td>0.04 ± 0.02</td>
</tr>
<tr>
<td>Stator iron (FE simulated)</td>
<td>0.0083</td>
<td>0.016</td>
</tr>
<tr>
<td>Magnet eddy-current (FE simulated)</td>
<td>0.006</td>
<td>0.012</td>
</tr>
<tr>
<td>Solid back-iron eddy-current (FE simulated)</td>
<td>-</td>
<td>1.2</td>
</tr>
<tr>
<td>Unaccounted</td>
<td>0.4</td>
<td>1.8</td>
</tr>
</tbody>
</table>

4.3.5 Eddy-Current Loss with Sintered and Bonded Magnets

The magnets used to construct the rotor (see Figure 4.9(b)) were sintered neodymium (1.5 x 10^{-6} \Omega m) with an order of magnitude lower resistivity compared to the bonded type (2 x 10^{-5} \Omega m). As a result, the eddy-current loss was predicted to be higher for sintered magnets. The FE simulated eddy-current losses in sintered and bonded magnets with solid back-iron under loaded conditions at 2,800 rpm are given in Figure 4.22 (see section 6.3.2 for the simulation procedures).

Figure 4.22. FE simulated magnet (1 pole) and total back-iron eddy-current loss waveforms for the test motor at 2,800 rpm with Rotor 1.
In the FE simulation, for simplicity only the magnet resistivity was changed and not the remanent flux density. In a practical design, the lower flux density of the bonded magnets will require changes to the magnet thickness and number of turns. Figure 4.22(a) gives the eddy-current loss versus rotor angle waveforms for both sintered and bonded magnets. The figure shows that the average loss values are 6.13 mW for sintered and 0.466 mW with bonded magnets. In addition, the eddy-current loss in the solid back-iron with resistivity \(2.22 \times 10^{-7} \Omega \text{m}\) was also simulated and it was found that the average value is 0.99 W (see Figure 4.22(b)).

![Figure 4.22](image)

(a) FE eddy-current loss of 1 pole magnet  
(b) FE total back-iron eddy-current loss

**Figure 4.23.** FE simulated magnet (1 pole) and total back-iron eddy-current loss at various speeds, no-load (NL) and load (L) conditions with Rotor 1.

Figure 4.23 gives the average magnet and back-iron eddy-current losses at different speeds under no-load (NL) and loaded (L) conditions. The plots indicate that eddy-current loss increases with the speed squared and it is not affected significantly by the load. Furthermore, the magnitude of such loss doubles with the presence of back-iron which is due to the increase in flux density (about 1.45 times) and this loss is proportional to the square of flux density (see Section 3.5).

The losses in the magnets are an order of magnitude lower than the copper losses (see Figure 4.26(a)), hence are negligible in this machine. As a result, the rotor with sintered magnets is preferred due to its better magnetic properties. Nevertheless, magnet losses can be more significant in larger machines and bonded magnets can be a possible solution in rotor design. This will be further investigated in Chapter 8.

From Figure 4.23(b), large eddy-current loss was predicted for the solid back-iron which is two orders of magnitude higher than loss in the magnets. This loss maybe
the dominant loss at high speed operation. A laminated AMM disk was constructed to be used as a back-iron but the design did not withstand the large axial magnetic forces, hence excluded.

### 4.4 Results - Motor Drive Tests

Figure 4.24 shows the FE simulated and directly measured phase voltage and current waveforms of the test motor drive without back-iron. The simulated results were generated with a time-stepping coupled-circuit. The simulation results are in good agreement with the measured waveforms. The small discrepancies in the waveforms may be due to small errors in the back-EMF magnitude and wave shape (see Figure 4.16(b)). In addition, the conduction losses of the power module were not considered in the simulation. Similar results were also observed with the back-iron configuration.

![FE and measured phase voltages](image1.jpg)

![FE and measured phase currents](image2.jpg)

**Figure 4.24.** FE simulated and measured phase voltage and current waveforms at 2,800 rpm with Rotor 1.

The FE simulated instantaneous torque waveform at 1,000 rpm is shown in Figure 4.25(a), which has a large torque ripple. As shown in the figure, the presence of the back-iron increases the torque. The average torque was obtained from the waveform captured and plotted together with the measured torque in Figure 4.25(b) with respect to speed. Figures 4.25(b) and 4.26 give the FE simulated torque, copper loss and efficiency plots at various speeds. It should be noted that since the input electrical power was not measured during the tests, it was estimated as the sum of measured output...
power, the copper loss from the measured input current, and the measured open-circuit loss at that speed. The “Measured” efficiency of the test motor was obtained using this estimated input power, which will be given later in Figure 4.26(b).

![Graph](image1)

(a) FE simulated torque, 1,000rpm

(b) FE and measured average torque

**Figure 4.25.** FE simulated and measured instantaneous and average output torques at 1mm airgap with Rotor 1.

![Graph](image2)

(a) FE and measured copper loss

(b) FE and measured efficiency

**Figure 4.26.** FE simulated and measured copper loss and efficiency at 1mm airgap with Rotor 1.

At 1,000rpm speed, the presence of the back-iron increases the predicted average torque by about 38% while decreasing the current by 16%. Even though the predicted iron loss was doubled with the back-iron configuration (see Section 4.3.4), it is still at least 20 times smaller than copper loss at this speed. In general, using the back-iron is predicted to reduce the input current and increase the average torque for the same operating speed. As a result, higher efficiency was predicted with back-iron and greater
torque can be obtained by increasing the current before reaching the thermal limit of the winding.

The measured results based on the without back-iron configuration were also included in Figure 4.26. The measured torque was about 20% and the copper loss about 60% less than the FE simulated values. Furthermore, the measured open-circuit loss (see Figure 4.21(b)) was an order of magnitude smaller than the copper loss. As a result, the measured efficiency was 6% higher than the FE simulated values. This showed that the copper loss was the dominant loss that influences the overall efficiency.

A high input current was drawn during testing with the back-iron configuration. This could be due to the high open-circuit loss. In order not to damage the motor windings, experimental measurements were not continued. Even though high efficiency was simulated, the high open-circuit loss was not taken into consideration and when this was included the efficiency was predicted to be lower than without the back-iron configuration (see Figure 4.27).

Figure 4.27 shows the contour plots of the calculated efficiency as a function of output torque and speed for both with and without back-iron configurations. Each contour plot was calculated based on a curve fit of the measured open-circuit loss characteristics (see Figure 4.28(a)) and the measured back-EMF and stator resistance. In this approach, it is assumed that the stator current is sinusoidal and in phase with the back-EMF waveform at all times.

![Figure 4.27](image)

**Figure 4.27.** Calculated efficiency contour map for the test machine at 1mm airgap with Rotor 1.
The gross electromagnetic torque was calculated as the sum of the mechanical output torque and the open-circuit power loss at that speed (see Equation 4.7). Then, the copper loss as a function of torque was found using Equation 4.8. It should be noted that the back-EMF constant \( k \) was obtained previously in Figure 4.18(a).

\[
T_{\text{Gross}} = T_M + T_{\text{OC}}(\omega_M) = \frac{3EI_{\text{ph}}}{\omega_M}
\]

\[
P_{\text{cu}} = 3I_{\text{ph}}^2R_{\text{wire}} = \frac{R_{\text{wire}}}{3k^2} \left[ T_M^2 + T_{\text{OC}}^2(\omega_M) \right]
\]

where \( T_{\text{Gross}} \): gross electromagnetic output torque, \( T_M \): mechanical output torque, \( T_{\text{OC}} \): open-circuit loss torque, \( I_{\text{ph}} \): phase current, \( \omega_M \): mechanical rotational speed and \( k \): phase induced back-EMF constant (mechanical).

The results were plotted in Figure 4.28(b) and the measured copper loss was about 16% higher than the estimated values at low torque (6.62 mNm). At higher torque values, the estimated loss were more comparable.

The points at which experimental efficiency measurements were taken are plotted as diamond dot points in the contour plot (see Figure 4.27(a)). The calculated efficiency characteristics at low torque values were about 9% higher than the measured results due to smaller copper loss prediction (see Figure 4.28(b)). At higher torque values, the calculated efficiency was about 2% higher as the copper loss prediction was more closer
to the measured loss. Also, the contour plots predicted about 11% lower maximum efficiency with the back-iron configuration over the speed and torque range considered, which is due to the higher open-circuit loss (see Figure 4.21(b)).

### 4.4.1 Magnet Size and Shape

Table 4.5 compares the experimental performance at similar values of output torque and speed of the test motor configurations with two pole rotors with neodymium magnets of different size and shape (see Figure 4.9). As it can be seen in the table, the three rotor types have significantly different efficiencies due to variations in the iron and copper losses. For example, Rotor 1 has the smallest back-EMF constant and so requires the highest current to generate the same level of output torque. This results in the highest copper loss and hence the lowest efficiency. On the other hand, Rotor 3 which has slightly larger magnets than Rotor 1, offers about 17% greater efficiency due to its higher back-EMF constant. In addition, Rotor 2 with the largest magnets has the highest back-EMF constant and hence the highest efficiency that was mainly due to the lower copper loss which was the dominant loss.

<table>
<thead>
<tr>
<th>Rotor</th>
<th>EMF constant</th>
<th>Output Power</th>
<th>Open-circuit Loss</th>
<th>Copper Loss</th>
<th>Efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.00931 V/rad/s</td>
<td>2.06 W</td>
<td>0.33 W</td>
<td>2.78 W</td>
<td>39.8 %</td>
</tr>
<tr>
<td>2</td>
<td>0.0160 V/rad/s</td>
<td>1.84 W</td>
<td>0.27 W</td>
<td>0.82 W</td>
<td>62.8 %</td>
</tr>
<tr>
<td>3</td>
<td>0.0145 V/rad/s</td>
<td>2.12 W</td>
<td>0.46 W</td>
<td>1.14 W</td>
<td>56.9 %</td>
</tr>
</tbody>
</table>

The open-circuit loss generally increases with the size of the magnet. This was demonstrated by comparing Rotor 3 and Rotor 1 which were constructed with disk shape magnets of 15 mm and 12 mm diameters. The increased loss was expected as the average tooth flux density increases with larger tooth area covered by the magnet. Also, as discussed previously, the higher resulting axial force increases the bearing loss which is a significant part of the open-circuit loss. Nevertheless, the shape of the magnets also have an important influence on the open-circuit losses as Rotor 2 with the largest magnets showed the lowest loss.
It should be noted that the disk and square magnets utilised in construction of the three rotors did not utilise the available rotor surface area well. Therefore, the designs considered in Chapter 5 utilised custom-built arc-shaped magnets for four and ten pole rotors of a 110\textit{mm} diameter AMM AFPM prototype.

### 4.4.2 Study of Airgap Length

The airgap length is an important parameter in axial-field PM machines as it affects the machine’s output torque and efficiency. In [71], the characteristics of a large airgap axial-field machine was investigated for a water pump application where the rotor and stator were sealed separately. The authors showed that the axial gap motor which had 80\% maximum efficiency and produce a torque of 0.27\textit{Nm} with a 8\textit{mm} airgap length was still able to maintain 66\% efficiency at a 12\textit{mm} airgap length with a torque of 0.12\textit{Nm}. Therefore to understand the effect of airgap length on AMM AFPM machines, the characteristics of the concept demonstrator machine at different airgap lengths are investigated.

In this study, measurements of the phase back-EMF voltage, open-circuit loss and efficiency were conducted under four different airgap lengths: 1\textit{mm}, 2\textit{mm}, 3\textit{mm} and 4\textit{mm}. The results given in Figures 4.29 and 4.30 were obtained using Rotor 2 which was chosen as it showed the highest efficiency.

![Figure 4.29](image)

**Figure 4.29.** Measured back-EMF and open-circuit loss at various airgap lengths with Rotor 2.

As expected, the phase back-EMF is linearly proportional to speed and decreases as the airgap length increases (see Figure 4.29(a)), and smaller airgap results in a higher back-EMF but higher open-circuit losses. In addition, the results show that open-circuit
loss increases with speed but decrease with airgap length. Nevertheless, for the test machine the open-circuit losses were an order of magnitude lower compared to copper loss that is given in Figure 4.30(a).

Figure 4.30 displays the test results of the measured copper loss and efficiency. Since the increase in the airgap length decreases the back-EMF and increases the no-load speed, higher current was required at the same operating speed. As a result, the copper loss also increases with airgap length as shown in Figure 4.30(a). In addition, as shown in Figure 4.30(b) that for a constant airgap length, increasing the output torque decreases the efficiency. This is because copper losses are dominant in this design, which are proportional to the square of the current while the output power is proportional to the current. This figure also shows that for a given torque, increasing the airgap length reduces the back-EMF which increases the required current and hence reduces the efficiency.

![Graphs showing measured copper loss and efficiency](image)

**Figure 4.30.** Measured copper loss and efficiency curves at various airgap lengths with Rotor 2.

Overall, the test results showed that the design achieved higher efficiency at small airgap length and the copper loss is the principal source of loss for operating speeds under 4,000 rpm.

### 4.4.3 Comparison of AMM and SMC based Test Motors

An SMC stator with identical dimensions and stator windings to the AMM stator was built for comparison. Both stators were tested using Rotor 2 (without back-iron) and the measured motor parameters are given in Table 4.6. It can be seen that the stators
have similar back-EMF constants, coil inductances and coil resistances. The AMM-based motor offers 45% less open-circuit iron loss than the SMC based motor. It was expected that the AMM stator should offer significantly lower iron loss than SMC by at least an order of magnitude (see Figure 3.7). The difference is associated with extra iron losses associated with the uncoated AMM ribbon and the high tension used to wind the core during construction to increase the stacking factor.

<table>
<thead>
<tr>
<th>Stator</th>
<th>EMF constant</th>
<th>Coil Inductance</th>
<th>Coil Resistance</th>
<th>Open-Circuit Iron Loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>AMM</td>
<td>0.0160</td>
<td>5.0</td>
<td>9.0</td>
<td>0.39</td>
</tr>
<tr>
<td>SMC</td>
<td>0.0158</td>
<td>4.7</td>
<td>8.9</td>
<td>0.88</td>
</tr>
</tbody>
</table>

The factors affecting the efficiency of the motors under load are summarised in the following two figures. Figure 4.31 shows the components of the power losses (copper and iron) and the measured output power. In Figure 4.32, the measured efficiency versus output torque characteristics are given.

![Figure 4.31](image-url)  
*Figure 4.31. Measured components of power characteristics of AMM and SMC motors (output power, open-circuit iron loss and copper loss) versus speed, at a DC link voltage of 15Vdc and an airgap of 1mm with Rotor 2.*
It can be observed from Figure 4.31 that the AMM-based motor develops slightly higher output power than the SMC motor. Given the similarity with the back-EMF constants, inductances and resistance, this difference is likely to be due to the open-circuit losses. In addition, as it can be seen in Figure 4.32, the efficiency of the AMM machine is also higher particularly at light loads due to its lower open-circuit losses which becomes close to the efficiency of SMC machine at higher output torques where the copper loss is dominant.

![Figure 4.32](image)

**Figure 4.32.** Measured efficiency versus output torque characteristics of AMM and SMC motors, at a DC link voltage of $15V_{dc}$ and airgap of an $1\text{mm}$ with Rotor 2.

Figure 4.33 shows contour plots of the calculated efficiency as a function of output torque and speed for both the AMM and SMC machines. The points at which experimental efficiency measurements were taken are also plotted as diamond dot points in the figures. The estimated efficiency characteristics were higher than the measured results (about 5% for AMM and about 7% for SMC) which is due to smaller predicted copper loss as discussed previously in Figure 4.28(b). In addition, the SMC machine has about 10% lower maximum efficiency compared to the AMM machine. Furthermore, the contour plots showed a smaller variation of efficiency at higher values of the output torque.
Figure 4.33. Calculated efficiency contour maps for AMM and SMC test machines with Rotor 2.
4.5 Conclusions

The research work in this chapter aimed to investigate the feasibility and characteristics of axial-field AMM machines based on a small concept demonstration stator. A 32mm diameter AFPM motor utilising slotted AMM and SMC stators was successfully designed and constructed. It was shown that the analytical design approach provides reasonable accuracy. The results were validated using a 3D FEA analysis tool and compared with experimental data obtained from a custom-built test setup. In addition, the performance characteristics of the motor were examined carefully and the results presented includes the motor efficiency contours.

The key results are:

- The results from the 3D finite-element analysis modeling tool utilised in this study showed good correspondence with the measured parameters of the AFPM machines. The prediction of axial force and the iron and eddy-current losses provided an in-depth understanding of the behaviour of the AMM motor.

- It was observed that the presence of rotor back-iron increases the airgap flux and hence increases the amplitude of both the back-EMF and open-circuit loss which have opposite effects on the efficiency.

- It was concluded that high open-circuit loss measured during the experiments were primarily due to the bearing and the rotor eddy-current losses. It was observed that the large axial force in the single-sided configuration increases the bearing loss and the solid rotor back-iron has high eddy-current losses (as simulated).

- The designed rotors utilised commercial available magnets without optimisation. The test results revealed that the magnet pole areas should cover the whole stator tooth’s surface area to maximise the back-EMF constant, which will be implemented in the later chapter.

- Larger airgap length reduces the open-circuit loss but also results in a lower back-EMF constant. For the motors considered, copper loss was found as the dominant factor. Therefore, it was concluded that increasing the airgap length decreases the efficiency due to the significant increase in input current.
• Another emerging magnetic material SMC is also utilised for comparison. The results showed that the AMM stator has a slightly higher torque and a significantly higher efficiency compared to the identical size SMC stator.

The large axial force, selection of the airgap length and open-circuit losses were further investigated in Chapters 7 and 8 with better designed and more practical large size (110mm) prototypes.
Chapter 5

110mm Machine Analytical Analysis and Design

In Chapter 4, a simple analytical approach was utilised to design the winding for a precut three-tooth concept demonstration small size (32mm) AMM stator based on two-pole configuration.

In this chapter, the analytical approach is further extended to design a larger size AMM stator for AFPM machine with concentrated stator windings and a surface-mounted magnet rotor. This includes determination of stator size, slot size, magnet thickness and slot and pole combination selection. Suitable analytical models from literatures are also utilised in the design and analysis. This includes analytical calculation of the machine’s parameters, open-circuit losses and performance.

In the design section, the analytical approach was implemented to investigate the design trade-off including AMM stator cutting cost. Then, the design process for the prototype was demonstrated based on the proposed guidelines. The stator has 12 slots (12S) which were cut from an AMM toroidal core sample of 110mm outer diameter provided by the industrial partner. Two motor designs were considered: a four-pole design (12S4P) and a ten-pole design (12S10P).

3D FEA was utilised to verify the analytical model based on the prototype model. The details of the 3D FE modeling will be discussed in Chapter 6.
5.1 Airgap Magnetic Flux Distribution

In analysing axial-field machines it is often useful to simplify it to an “equivalent” radial-field or linear machine to allow the standard analysis methods to be used. A common approach is to analyse the AFPM stator tooth and rotor magnet topology at a given radius. As this topology changes as a function of radius, often the average radius is used.

The analysis of electrical machines often begins with determining the airgap flux distribution under open-circuit conditions. This then allows the stator tooth and yoke flux densities to be found. From these densities the back-EMF waveform and open-circuit losses can be obtained.

Several analytical models were developed in [79, 127–129] to calculate the airgap field distribution for surface permanent magnet radial-field machines. In these studies, the airgap magnet field was derived by assuming an infinite permeability in the stator and rotor back-iron. In addition, the analytical expressions of the airgap field distribution initially neglect slotting in the stator and the slotting effect can be accounted later using a permeance function.

Similarly, several analytical models were also developed in [130–138] specifically for axial-field machines to obtain the open-circuit flux density distribution in the airgap region. In [139], the flux density within the machine core was derived. The model in [138] was implemented in this work where the axial-field machine at its average radius was modelled as a linear machine and the field distribution expressed in terms of Cartesian co-ordinates. In the model, the stator and rotor are assumed to be smooth and have infinite permeability. By solving the 2D Poisson’s equation for the magnetic field distribution, it can be shown that the magnetic field in the airgap at the surface of the stator teeth is given by Equations 5.1 to 5.3.

\[
B_{gPM}(x) = - \sum_{n=1,3,5,\ldots}^{\infty} \frac{B_{PM1}}{B_{PM2}}
\]

\[
B_{PM1} = \frac{8B_r}{n\pi} \sin \left( \frac{\alpha_m n \pi}{2} \right) e^{-n\pi \xi / \tau_{pole}} \cos \left( \frac{n\pi x}{\tau_{pole}} \right)
\]

\[
B_{PM2} = \left( e^{2n\pi \xi / \tau_{pole}} + 1 \right) \frac{\mu_r \left( -e^{-2n\pi \xi / \tau_{pole}} + 1 \right) \left( e^{2n\pi \xi / \tau_{pole}} + 1 \right)}{\mu_0 \left( e^{2n\pi \xi / \tau_{pole}} - 1 \right)}
\]
where $\tau_{pole}$: pole pitch, $\alpha_m$: ratio of magnet width to pole pitch and $\mu_r$: relative recoil permeability of the magnet.

In [140–143], Zhu and Howe introduced models to consider the slotting effects in electrical machines. These were based on modelling the fringing flux in the stator slot openings as having a semi-circular flux path from the airgap to the sides of the stator teeth where the centre of the path is the corner of the tooth (see Figure 5.1). The relative permeance function $\tilde{\lambda}$ in Cartesian coordinates presented in [140] is adapted in this research. The permeance $\lambda$ and relative permeance $\tilde{\lambda}$ can be calculated from Equations 5.4 and 5.5.

\[
\lambda = \frac{\mu_0}{I_g + \frac{I_m}{\mu_r} + \frac{2\pi r_s}{4}} \quad (5.4)
\]

\[
\tilde{\lambda}(x) = \frac{\lambda}{\left(\frac{\mu_0}{I_g + \frac{I_m}{\mu_r}}\right)} \quad (5.5)
\]

where $r_s$: radius of fringing flux path, which is non-zero over stator slots but zero over stator teeth.

The airgap flux distribution for a slotted core $B_{gSlot}$ can be calculated as in Equation 5.6 using the relative permeance function $\tilde{\lambda}(x)$. In addition, the tooth flux $\Phi(x)$ can be obtained as in Equation 5.7 by integrating numerically the airgap flux density distribution.

\[
B_{gSlot}(x) = \tilde{\lambda}(x)B_{gPM}(x) \quad (5.6)
\]

\[
\Phi(x) = \int_0^{\tau_{pole}} B_{gSlot}(x)dx \quad (5.7)
\]
3D FEA was utilised to verify the analytical prediction using the AMM stator structure considered in this research. Figures 5.2 and Table 5.1 give the dimensions of the $110\,mm$ diameter, parallel-slot axial-field stator modelled in the finite element analysis which forms the reference model for the calculation. Note that the details of the choice of the dimensions will be given later in Section 5.11.

![Diagram of reference machine R1 stator dimensions](image)

**Figure 5.2.** Reference machine R1 stator dimensions.

<table>
<thead>
<tr>
<th><strong>Stator Dimensions (mm)</strong></th>
<th><strong>Parameters (mm)</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>Outer Diameter (OD) 110</td>
<td>Magnet Thickness ($l_m$) 3</td>
</tr>
<tr>
<td>Inner Diameter (ID) 45</td>
<td>Airgap Length ($l_g$) 1</td>
</tr>
<tr>
<td>Slot Depth (SD) 12</td>
<td></td>
</tr>
<tr>
<td>Slot Width (SW) 8</td>
<td></td>
</tr>
<tr>
<td>Yoke Thickness (YT) 18</td>
<td></td>
</tr>
</tbody>
</table>

**Table 5.1.** Reference machine R1 dimensions.

Figures 5.3 and 5.4 compare the airgap flux density plots of four and ten-pole designs for the above machine at the average radius for non-slotted and slotted cores that are calculated based on the analytical equations and FE simulations. As can be seen from the figures, the waveforms were similar with small discrepancies particularly over the slot openings.
Figure 5.3. Average radius airgap flux density plots of non-slotted and slotted cores with 12 slot and 4 pole (12S4P).

Figure 5.4. Average radius airgap flux density plots of non-slotted and slotted cores with 12 slot and 10 pole (12S10P).

5.2 Tooth Magnetic Flux Density

As it is known, the slotted stator provides a strong support for the windings and it allows room for the coils while maintaining a small effective airgap. Therefore, this structure is adopted in this research and will be analysed in details.

In a slotted axial-field machine, it was assumed that the airgap flux travels axially from the rotor pole magnet enters the stator teeth and passes circumferentially along
the yoke and then returns axially back out of the stator teeth. For simplicity of the analytical approach, it is assumed that there is no flux movement in the radial direction in the stator. This is a reasonable assumption considering that the stator consists of a the tape wound toroidal ring core and hence there is low permeability in the radial direction. In addition, it was shown in [121, 122] that although there is a degree of radial flux in axial-field design it is significantly smaller than the main flux. This is also verified in this research in Section 5.2.1 using the finite element modelling.

It should be noted that the zero radial flux assumption means that the highest yoke flux density will occur at the outer diameter of the machine. This can be deduced from the fact that the circumferential magnet length is largest at the outer radius and can provide the greatest driving flux into the stator back-iron.

![Figure 5.5. A section of a slotted AMM stator core showing tooth dimensions.](image)

As the flux distribution tends to concentrate in the stator teeth due to fringing effects, the tooth flux density is expected to be higher than the average airgap flux density. It should be noted that the stator tooth width in axial-field design varies as a function of radius as shown in Figure 5.5. Furthermore, assuming a constant amount of fringing flux, the stator tooth flux density would be expected to be largest at the inner radius and smallest at the outer radius. The shape of the permanent magnet also affects the flux densities in the tooth and yoke as demonstrated in [144].

In this research, the analytical approach for modelling the tooth flux density focuses on arc-shaped magnets and uses two simple analytical models based on Model B1 and Model B2 which will be explained in the following sections and compared to 3D finite element simulations.
5.2.1 Model B1 and Model B2

In Model B1, the airgap flux density is assumed to be uniform over the stator teeth and is assumed zero over the slots by neglecting fringing (see Figure 5.6).

\[ B_{B1} \approx B_g \left( \frac{l_m}{l_{mr}} \right) \left( \frac{l_m}{l_{mr}} + l_g \right) \]  

(5.8)

Figure 5.6. Assumptions in tooth flux density of Model B1.

For a non-slotted core, the airgap magnetic flux density \( B_g \) produced by a permanent magnet with a linear demagnetisation curve can be approximated by Equation 5.8 [145]. The model approximated tooth flux density \( B_{tB1} \) is equivalent to \( B_g \) which was the model utilised in Chapter 4.

In Model B2, a uniform (non-slotted) stator is assumed and the airgap flux density magnitude thus uniform over all the airgap (see Figure 5.7). Carter’s coefficient is utilised in the model to account for the slotting effects in calculating the average airgap flux density.

\[ B_{tB2} \approx B_g \left( \frac{l_m}{l_{mr}} \right) \left( \frac{l_m}{l_{mr}} + l_g \right) \]  

(5.9)

Figure 5.7. Assumptions in tooth flux density of Model B2.

Since, the airgap flux density decreases over slot openings, in order to take the slotting into account, a correction coefficient can be included which models the lower mean...
magnetic flux density as fictitious increase in the airgap. As a result, the effective electromagnetic airgap $l'_{em}$ is approximated using the Carter’s coefficient $k_c$ and actual electromagnetic airgap $l_{em}$ (see Equation 5.13). According to [141], the Carter’s coefficient can be calculated from Equation 5.9-5.12 where $k_c \geq 1$. For slotless designs the Carter’s coefficient $k_c = 1$. In addition, the slotted average airgap flux density can then be calculated using Equation 5.14.

\[
l_{em} = l_g + \frac{l_m}{\mu_r} \tag{5.9}\]

\[
k_c = \frac{\tau_{slot}}{\tau_{slot} - \gamma_{c} l_{em}} \tag{5.10}\]

\[
\tau_{slot} = \Gamma_{tooth} + \Gamma_{slot} \tag{5.11}\]

\[
\gamma_{c} = \frac{4}{\pi} \left[ \left( \frac{\Gamma_{slot}}{2l_{em}} \right) \arctan \left( \frac{\Gamma_{slot}}{2l_{em}} \right) - \ln \left( \sqrt{1 + \left( \frac{\Gamma_{slot}}{2l_{em}} \right)^2} \right) \right] \tag{5.12}\]

\[
l'_{em} = k_c l_{em} \tag{5.13}\]

\[
B_{TB2} = \frac{B_r \left( \frac{l_m}{\mu_r} \right)}{l'_{em}} \tag{5.14}\]

where $\tau_{slot}$: slot pitch, $\Gamma_{tooth}$: arc length of stator tooth, $\Gamma_{slot}$: arc length of stator slot and $B_{TB2}$: average airgap flux density.

As shown in Figure 5.5, the stator tooth width and tooth pitch varies as a function of radius for parallel slot stators. As a result, assuming that the magnet pole arc is 180° electrical then the tooth flux density also changes along the radius. A first approximation of the stator tooth flux density is given by the average airgap flux density divided by the ratio of stator tooth width to stator tooth pitch at the respective radius. Hence, the tooth flux density $B_t$ at radius can be given by

\[
B_t(r) = B_{TB2}(r) \left( \frac{\Gamma_{tooth}(r) + \Gamma_{slot}(r)}{\Gamma_{tooth}(r)} \right)
\]

\[
= B_{TB2}(r) \left( \frac{\Theta_{tooth}(r) + \Theta_{slot}(r)}{\Theta_{tooth}(r)} \right) \tag{5.15}\]

where all the parameters are given as a function of radius $r$, $\Theta_{slot}$: arc angle of slot and $\Theta_{tooth}$: arc angle of tooth.
Note that for the parallel slot stator, the slot width $\Gamma_{\text{slot}}$ is constant and the slot angle is a function of radius $r$. Therefore, the tooth angle $\Theta_{\text{tooth}}$ is calculated from the slot angle and the stator tooth pitch are given in Equations 5.16 and 5.17. The tooth flux density at the stator inner, average and outer radius can be calculated using these equations.

$$\Theta_{\text{slot}}(r) = \frac{\Gamma_{\text{slot}}}{r} \tag{5.16}$$

$$\Theta_{\text{tooth}}(r) = \frac{2\pi Q_s}{Q_s} - \Theta_{\text{slot}}(r) \tag{5.17}$$

where $Q_s$: number of slots.

The analytical calculations were compared with FE simulations. The dimensions of the machine design analysed in the calculations and modelled in the finite element was given earlier in Table 5.1. The two models described above utilise this machine and consider only the axial flux and neglect any flux in the radial direction. To test this assumption, Figures 5.8(a) and 5.8(b) are given where radial flux density versus angular position waveforms are simulated by FE modelling at the average radius within the stator yoke for isotropic (100% stacking factor) and anisotropic (90% stacking factor) AMM toroids. In addition, the axial flux density versus angular position at an axial position corresponding to the surface of the teeth at various radii are also given in Figures 5.8(c) and 5.8(d).

It can be seen in the figures that comparing the amplitudes of the radial and axial waveforms, the radial flux of the four-pole stator is much smaller, about 30% for the isotropic and 5% for the anisotropic materials. The percentages are even smaller for the ten-pole stator. Therefore, it can be concluded that the assumption to neglect the radial flux in the models are a good approximation.

Table 5.2 summarises the analytical values based on the models and FE simulated results in Figures 5.8(c) and 5.8(d). As it can be seen, Model B1 predicted about 16% higher airgap values to the FE values while Model B2 with 3%. In addition, Model B1 showed about 22% smaller tooth flux density values at average radius compared to the FE values. For Model B2, about 10-20% discrepancies for density values at average and outer radius were obtained. However, the calculated values of Model B2 at inner radius was twice the value of FE result which is believed to be due to high leakage flux with the much smaller tooth arc length.

Therefore, Model B2 was found to have the best accuracy in these studies (although it overestimates the flux density).
Figure 5.8. FE simulated radial flux density waveforms versus angular position at the average radius within the stator yoke, and axial flux density waveforms versus angular position at axial position corresponding to the tooth tip at various radii for 4-pole (4P) and 10-pole (10P) machines.

Table 5.2. Summary results of the analytical (Model B1 and Model B2) and finite element (FE) based model.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Analytical Model B1</th>
<th>Analytical Model B2</th>
<th>FE</th>
</tr>
</thead>
<tbody>
<tr>
<td>12S4P/12S10P</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Peak airgap $B_g$ (T)</td>
<td>0.93</td>
<td>-</td>
<td>0.80</td>
</tr>
<tr>
<td>Mean airgap flux density $B_g$ (T)</td>
<td>-</td>
<td>0.83</td>
<td>0.80</td>
</tr>
<tr>
<td>Peak tooth inner radius $B_{Tin}$ (T)</td>
<td>0.93</td>
<td>2.34</td>
<td>1.17</td>
</tr>
<tr>
<td>Peak tooth average radius $B_{Tave}$ (T)</td>
<td>0.93</td>
<td>1.36</td>
<td>1.13</td>
</tr>
<tr>
<td>Peak tooth outer radius $B_{Tout}$ (T)</td>
<td>0.93</td>
<td>1.18</td>
<td>1.06</td>
</tr>
</tbody>
</table>
5.2.2 Tooth Magnetic Flux Waveform

From the airgap flux density distribution, the tooth flux waveform and hence the phase back-EMF waveform can be estimated using the sliding array method described in Section 4.1.1. The analytically generated and FE simulated phase flux-linkage and back-EMF plots are shown in Figure 5.9 based on analytical Model B1 at the average radius.

![Figure 5.9](image)

(a) Phase flux-linkage  
(b) Phase back-EMF

**Figure 5.9.** Analytical and FE simulated phase flux-linkage and back-EMF plots for 12 slot and 4 pole (12S4P) test machine at 1,000rpm.

As shown in Figure 5.9(a), the peaks have similar flat-top but smoother edges and higher amplitudes were observed for the FE simulated waveform. This is due to the analytical Model B1 neglecting fringing effects and also because the airgap flux distribution was calculated only at the average radius and did not take into account the variation of this distribution with radial position. However, the analytical method can still generate the general shape of the back-EMF waveform.
5.3 Stator and Rotor Back-Iron Magnetic Flux Densities

To estimate the magnetic flux density in stator yoke, the amount of flux traveling axially down the stator teeth produced by each rotor magnet pole needs to be calculated first. A rough approximation to the rotor pole flux \( \phi_p \) can be estimated using Equation 5.18 below as the product of the magnet pole area and the peak airgap flux density \( B_g \).

\[
\phi_p = \left( \frac{\Theta_{Mag}}{2} \right) (R_{MagOut}^2 - R_{MagIn}^2)B_g
\]

where \( \Theta_{Mag} \): arc angle of magnet pole, \( R_{MagOut} \): magnet outer radius and \( R_{MagIn} \): magnet inner radius.

Two stator yoke models (Model By1 and Model By2) are proposed. Model By1 neglects the effect of the stator slotting. Model By2 considers Carter coefficient that results in a lower value for \( B_g \).

In the stator back-iron, the rotor pole flux splits into two circumferential paths. Hence, the stator back-iron magnetic flux density \( B_y \) is approximated with the rotor pole flux divided by two and then divided by the stator back-iron cross-section as given in Equation 5.19. This equation assumes the stator flux is uniformly distributed over the stator yoke and hence assumes the presence of some radial flux movement in the yoke.

\[
B_y = \frac{\left( \frac{\phi_p}{2} \right)}{(R_{StaOut} - R_{StaIn})h_{Sta}}
\]

where \( h_{Sta} \): stator yoke thickness.

A rough estimation of the average magnetic flux density in the rotor back-iron \( B_{RBI} \) can also be calculated using the method described above and is given below in Equation 5.20.

\[
B_{RBI} = \frac{\left( \frac{\phi_p}{2} \right)}{(R_{RotOut} - R_{RotIn})h_{Rot}}
\]

where \( R_{RotOut} \): outer radius of the rotor, \( R_{RotIn} \): inner radius of the rotor and \( h_{Rot} \): rotor back-iron thickness.

Table 5.3 is provided to summarise the analytical and FE results of the peak flux density in the stator and rotor back-iron for the four-pole and ten-pole machines. The FE
calculation was performed assuming a solid (unlaminated) yoke. It should be noted that increasing the pole numbers reduce the flux per pole and hence the back-iron flux densities by a factor of about 1.4, which is visible in the stator yoke flux density. Although, the rotor yoke flux density does not change significantly this is due to different thickness rotor back-iron in the test motors to avoid excessive saturation.

Table 5.3. Analytical and FE results of stator and rotor back-iron flux densities under open-circuit conditions of two machines.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Analytical Model By1</th>
<th>Analytical Model By2</th>
<th>FE Simulation</th>
</tr>
</thead>
<tbody>
<tr>
<td>12 slot 10 pole machine (12S10P)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Stator yoke $B_y(T)$</td>
<td>0.52</td>
<td>0.47</td>
<td>0.41</td>
</tr>
<tr>
<td>Rotor back-iron $B_{RBI}(T)$</td>
<td>0.86</td>
<td>0.76</td>
<td>0.78</td>
</tr>
<tr>
<td>12 slot 4 pole machine (12S4P)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Stator yoke $B_y(T)$</td>
<td>1.47</td>
<td>1.30</td>
<td>1.18</td>
</tr>
<tr>
<td>Rotor back-iron $B_{RBI}(T)$</td>
<td>1.37</td>
<td>1.21</td>
<td>1.24</td>
</tr>
</tbody>
</table>

As can be observed in the table, the calculated stator back-iron flux density is about 25% higher for Model By1 and 12% for Model By2. On the other hand, the calculated rotor back-iron flux density is about 10% higher for Model By1 and 3% lower for Model By2. Therefore, Model By2 is found to have better accuracy.

5.4 Machine Sizing Based on Shear Stress

The airgap shear stress ($N/m^2$) which is the tangential force on the conductors in each stator slot per unit airgap length per unit stack length, provides useful performance metrics (i.e. output torque) for electrical machines. This shear stress expression also can be written as the product of the magnetic and electric loading, which are key concepts in designing electric machines.

Based on the concepts of shear stress and magnetic and electric loading, the stator outer diameter and optimum inner to outer diameter ratio of the slotted AFPM machine are to be investigated in this section. This includes derivation of output torque expression for slotted core and selection of airgap length and magnet thickness based on the outer diameter.
5.4.1 Magnetic and Electric Loading

Magnetic and electric loading provide a good insight of how the machine’s design parameters can affect the output torque. Fundamentally electric loading is limited by the temperature rise of the stator coils and magnetic loading by saturation of the magnetic materials.

The allowable magnetic loading is determined by the saturation in the teeth and yoke of the stator core (refer Section 5.2 and 5.3). In an axial-field machine, stator teeth saturation is likely to be worst at the inner radius due to the reduced tooth width but constant amount of fringing flux. From the stator teeth, the magnet flux then passes circumferentially along the yoke before returning axially back through the stator teeth. The stator teeth and yoke are required to be wide and thick enough so not to saturate excessively. Further investigation on the effect of saturation on the output torque will be described later in Section 6.4.2.

The electric loading is the linear current density in the machine. The key variables affecting the electric loading are the ratio of the slot width to slot pitch, the slot depth, the slot packing factor and the allowable conductor current density.

In a practical machine design, there is a trade-off between the magnetic and electric loading, taking into account other aspects such as thermal performance, efficiency and cost. The general trade-off is that:

- maximising the magnetic loading usually involves minimising the slot width to minimise stator teeth saturation, and maximising the stator yoke thickness (minimising the slot depth) to minimise stator yoke saturation.

- maximising the electric loading usually involves increasing the slot volume by maximising the slot width and slot depth.

In addition, the minimum allowable slot width depends on the inner diameter, the electrical loading and the saturation flux density of the stator teeth. On the other hand, the slot depth is determined by magnetic loading which is the saturation value in the stator yoke.
5.4.2 Torque Calculation

The force on a single stator wire can be calculated from the fundamental force equation \( F = BIL \) which is the interaction between magnetic flux density \( B \) with the wire of \( l \) length and carrying constant current of \( I \). The average airgap flux density \( B_{\text{ave}} \) (magnetic loading) can be found using analytical Model B1 as given below in Equation 5.21.

\[
B_{\text{ave}} = \left( \frac{2\pi r - Q_s \Gamma_{\text{slot}}}{2\pi r} \right) B_g \tag{5.21}
\]

The torque on the wire with a torque radius of \( r \) metres can be calculated as \( T = Fr \), where \( F \) is the force. For a slotted core with \( N_{\text{wires}} \) number of turns, the length \( l \) is defined as the active radial length of a conductor which is the stack length \( \Gamma_{\text{StackL}} \) and \( r \) is the average radius \( R_{\text{ave}} \) of the stator.

\[
T = B_{\text{ave}} l N_{\text{wire}} Ir \tag{5.22}
\]

The value \( N_{\text{wire}} \times I \) in the above equation can be calculated from the number of slots \( Q_s \), the cross-sectional area of each slot \( A_{\text{slot}} \), the slot copper packing factor \( p_{f_{\text{slot}}} \) (usually 20-60\%) and the allowable copper current density \( J \) (see Section 5.8).

\[
T = B_{\text{ave}} (\Gamma_{\text{StackL}})(Q_s A_{\text{slot}} p_{f_{\text{slot}}} J)r \tag{5.23}
\]

Therefore, first approximation of the output torque based on calculations at average radius can be given by

\[
T = B_{\text{ave}} (\Gamma_{\text{StackL}})(Q_s \Gamma_{\text{slot}} \Gamma_{\text{depth}} p_{f_{\text{slot}}} J) R_{\text{ave}} \tag{5.24}
\]

where \( \Gamma_{\text{depth}} \): slot depth.

The stator tooth width and tooth pitch for parallel slot stators varies with radial position and hence the average airgap flux density also changes along the radius. Therefore, the torque equation can be written using two parameters: the stator outer diameter and the ratio of inner to outer diameters.

5.4.3 Optimum Inner/Outer Radius

In rotating electrical machine design, the output torque is an important requirement. The size of the machine is one of the main factors determining the maximum torque
capability of a machine. According to [111, 112, 144], the output torque increases with the cube of outer diameter of the non-slotted stator core. In addition, the output torque can be expressed in terms of the ratio of inner to outer diameters which is an important design parameter for an AFPM machine. The optimal diameter ratio for an idealised non-slotted AFPM machine of 0.58 and a practical optimal value between 0.6 to 0.7 were derived in [111, 144]. In the analysis, a constant average magnetic flux density (magnetic loading) and a constant value of linear current density (electric loading) at the inner radius was assumed. The relationship of the output torque with the inner/outer radius for a slotted stator core is investigated in this section.

In a practical slotted core machine with a given stator outside diameter, the assumption of constant linear current density at the inside diameter has limited validity particularly at the limiting case where the diameter ratio approaches zero. Therefore, a more realistic model is to assume a constant value of current density. In this work, the linear current density $A(r)$ (electric loading) at radius $r$ is defined as in Equation 5.25.

$$A(r) = \frac{mN_{ph}I}{2\pi r}$$  \hspace{1cm} (5.25)

where $m$: number of phases and $N_{ph}$: number of phase turns.

The elementary torque component $\Delta T$ on radius $r$ (see Figure 5.10) can be obtained as in Equation 5.26 with the airgap flux density $B_{ave}(r)$ (see Equation 5.21).

$$\Delta T = \Delta Fr = [B_{ave}(r)A(r)2\pi r\Delta r]r$$  \hspace{1cm} (5.26)
The total machine electromagnetic torque $T$ can be calculated by integrating the elementary torque component $\Delta T$ between the inner and outer radii as illustrated in Figure 5.10. As it can be seen, the linear current density is highest at the inner radius of the stator core $R_{StaIn}$ for an axial-field structure. Therefore, the linear current density at the inner radius is defined in the elementary torque component $\Delta T$. Then the total electromagnetic torque can be given by Equation 5.27.

Note that, in order to examine the influence of the outer diameter, the total torque is expressed in terms of the outer radius ($R_{StaOut}$) and the ratio of the inner and outer radii defined as $k_r$ (see Equation 5.28). Substituting the definitions into the elementary torque component in Equation 5.26 and integrating the Equation 5.27, the total torque can be obtained as in Equation 5.29.

$$T = \int_{R_{StaIn}}^{R_{StaOut}} \left[ \left( \frac{2\pi r - Q_s \Gamma_{slot}}{2\pi r} \right) B_g \left( A(R_{StaIn}) \frac{R_{StaIn}}{r} \right) (2\pi r) r \right] dr \quad (5.27)$$

$$k_r = \frac{R_{StaIn}}{R_{StaOut}} \quad (5.28)$$

$$T = \pi B_g A(R_{StaIn}) k_r R_{StaOut}^2 \left( \pi R_{StaOut} (1 - k_r^2) - Q_s \Gamma_{slot} (1 - k_r) \right) \quad (5.29)$$

Note that, a torque equation for a non-slotted core (as it was derived in [144]) also can be obtained by using a zero number of slots $Q_s$ in the above equation.

The assumption of constant linear current density at the inner radius can be removed using Equation 5.30 based on a similar magnetomotive force per slot $N_{ph} I$ definition as in Equation 5.23.

$$A(r) = \frac{Q_s \Gamma_{slot} \Gamma_{depth} P_f I}{2\pi r} \quad (5.30)$$

Then, the electromagnetic torque equation produced is given in Equation 5.31. This equation can be rewritten as in Equation 5.32, where it was assumed that the parameter $Q_s \Gamma_{slot} / (2\pi R_{StaOut})$ which is the product of the number of slots times the slot width divided by the stator outer circumference, is a constant.

$$T = B_g \left( \frac{Q_s \Gamma_{slot} \Gamma_{depth} P_f I}{2\pi R_{StaIn}} \right) k_r R_{StaOut}^2 \left( \pi R_{StaOut} (1 - k_r^2) - Q_s \Gamma_{slot} (1 - k_r) \right) \quad (5.31)$$

$$T = B_g \left( \frac{Q_s \Gamma_{slot}}{2\pi R_{StaOut}} \right) \left( \Gamma_{depth} P_f I \right) R_{StaOut}^2 \left( \pi R_{StaOut} (1 - k_r^2) - Q_s \Gamma_{slot} (1 - k_r) \right) \quad (5.32)$$
The torque production capability of the axial machine with parallel slots based on the design model given in Table 5.1 and in Equation 5.32 as a function of diameter ratio $k_r$ is plotted in Figure 5.11. Note that, the outer diameter was kept constant in the plot and the inner diameter was varied according to the diameter ratio, and the graph was scaled for the maximum torque to be equal to unity.

![Figure 5.11. Analytical calculated electromagnetic torque of the reference axial-flux machine design as a function of the diameter ratio.](image)

It can be seen in the above figure that the torque increases with decreasing diameter ratio and reaches a peak value at a particular value of diameter ratio (minimum diameter ratio $k_{rmin}$). Two cases are considered for the torque curve in the region less than $k_{rmin}$. For Case a, the slot width was kept constant while the slot width was varied along with the inner diameter of the core for Case B. Figure 5.12 shows drawings of the cores. The torque reaches a peak value at $k_{rmin}$ and remained constant since the size of the teeth was not changed as the inner diameter was reduced in Case a. On the other hand, the torque dropped after the peak value at $k_{rmin}$ for Case B due to the decreasing slot width which also reduces the electric loading. For a slotted core, the region below $k_{rmin}$ has limited practicality and hence will not be included in the following analysis on the effect of changing the slot depth and slot width.
Assuming narrow slots, the minimum diameter ratio $k_{rmin}$ can be given in Equation 5.33 which is an important machine design parameter. For the reference model, the minimum diameter ratio $k_{rmin}$ is 0.28 as marked in Figure 5.11. In addition, a minimum stator inner diameter is required to have a reasonable rotor shaft diameter for the design speed range after allowing for the radial thickness of the end-windings.

$$k_{rmin} = \frac{Q_s \Gamma_{slot}}{2\pi R_{StaOut}}$$  \hspace{1cm} (5.33)

Figure 5.13 is provided to examine the effect of number of slots, slot depth and slot width on the torque versus diameter ratio curves. Note that, these curves have also been normalised to the maximum output torque of the reference design (12 slots, 12mm slot depth, 8mm slot width).

As it can be seen in the figure, the torque increases with the electric loading as the number of slots or the slot depth are increased which scales the curves vertically, and the magnetic loading is unchanged with the constant slot width. On the other hand, increasing the slot width decreases the magnetic loading while increasing the electric loading which changes the maximum torque (see Figures 5.13(b) and 5.13(d)). Furthermore, it can be shown using Equations 5.32 and 5.33 that maximum output torque is obtained at $k_{rmin} = 0.36$. For the 12 slot design this corresponds to a slot width of 10mm.
while for the 8 slot design it corresponds to a slot width of 15\,mm. In addition, for the reference model with slot width 12\,mm and a diameter ratio of 0.42 and comparable torque magnitude can be achieved using a stator with 8 slots and a diameter ratio of 0.28.

Figure 5.13. Analytical calculated electromagnetic torque of the reference axial-flux machine design as a function of the diameter ratio for various slot depths and widths. The torque is normalised against the maximum torque for the 12 slot machine with a slot depth of 12\,mm and a slot width of 8\,mm.
Figure 5.14 shows the stator iron volume comparison of 12 and 8 slot models with 12\text{mm} slot depth and 8\text{mm} slot width. The values of the diameter ratios required to produce the same output torque for each design are also plotted.

![Graph showing stator iron volume versus diameter ratio for 12 and 8 slot designs with slot depth 12\text{mm} and slot width 8\text{mm}.](image)

**Figure 5.14.** Analytical calculated stator iron volume versus diameter ratio for 12 and 8 slot designs with slot depth 12\text{mm} and slot width 8\text{mm}.

It can be seen in the above figure that 8 slot design has a higher stator iron volume which increases both the material cost and the iron loss. However, the increased iron loss in 8 slot design might not be as critical compared to the cutting cost for 12 slot design. For example, the AMM material considered in this work offers very low iron loss but the cutting and handling forms a significant fraction of the total manufacturing cost. The AMM cutting technique is discussed briefly in Section 5.11.1 where it is explained that the cutting cost is affected by both the number of slots but also by the length of the slots. Therefore, it should be considered that the 8 slot design has fewer slots but it has a smaller diameter ratio and hence longer slots. A more detailed analysis of the cutting process is required for a detail comparison but this is not part of the scope of this thesis.
5.4.4 Airgap Length and Magnet Thickness

In electrical machines, the maximum overall axial length of the machine depends on the applications, and the airgap length and magnet thickness affect the overall axial length, airgap flux density, back-iron thickness, type of bearing and shaft design. In addition, the selection of the airgap is affected by mechanical constraints due to the large axial attractive forces between the rotor magnets and the stator core in the axial structure. On the other hand, both the magnetic loading and cost depend on the thickness of the magnets. Excessive magnet thickness should be avoided as the magnet cost can be significant. In addition, the larger axial force and larger magnet may require extra care in handling and construction which can increase manufacturing cost significantly.

Neodymium magnets are known for their high remanent magnetic flux density. In order to achieve the required flux density in the machine, the required magnet thickness, volume and weight are usually small. In addition, the price of rare earth magnets has dropped substantially recently. Hence, neodymium magnets are popular in permanent magnet machines. In this work, only ring and arc type magnets are considered as they are better suited to the parallel slot stator. The eddy-current loss in sintered magnets (low resistivity) can be significant especially for fractional-slot designs. Therefore, a design using bonded magnets (with high resistivity) was also examined in this research.

As it is discussed in [15,83], the length of the air-gap ($l_g$) and magnet thickness ($l_m$) can be related to the stator core outer diameter ($D_o$) by

$$l_g = 0.006(D_o)$$  \hspace{1cm} (5.34)

$$l_m = \frac{4}{3}l_g$$  \hspace{1cm} (5.35)

Note that, the thickness of the magnet should be calculated with the effective airgap augmented to take into account of the slotting effects (see Section 5.2.1) and the finite permeability in stator core.

Figure 5.15 gives the calculated airgap length and magnet thickness as a function of outer diameter based on the above equations. The actual values used in some prototype machines reported in literature are also plotted in the figure. The actual magnet thicknesses were generally higher than the calculated values which is not surprising.
due to the reasons discussed previously. The choice of the magnet thickness also de-
pends on the required airgap flux density and cost as mentioned above. However, the
above equations can provide a starting point at the initial design stage.

![Figure 5.15. Airgap length and magnet thickness as a function of the core outer diameter.](image)

### 5.5  Stator Winding Design

The winding design in this research work consists steps to determine the approximate
number of turns and size of the copper wires based on the induced back-EMF equation
and available slot size. In order to estimate the induced back-EMF and slot size, the
winding factor is first calculated and the slot and pole combination is selected. The
following sections describe the procedure in details.

#### 5.5.1  Winding Factor Calculation

The induced voltage is affected by the distribution and pitching of the stator coils.
Therefore, the winding factor ($k_{w1}$) of the main harmonic (fundamental) which is the
product of the distribution factor and the pitch factor is included in the induced volt-
age equation as will be shown later in Equation 5.40. In addition, the fundamental
harmonic of order 1 for a two-pole fractional-slot winding machine was used in this
study so that all the sub-harmonics are taken into account [75].
In general, the winding factor can be calculated from the phase EMF phasor vector graph (EMF method) or by using analytical equations. In addition, specific analytical equations are required for different winding types. The two main concentrated winding types are double-layer designs where there is a coil around every tooth, and single-layer designs where there is a coil around every second tooth. This work focuses on double-layer concentrated windings.

The EMF method can be used for both single or double-layer concentrated windings. According to [83], the EMF phasor for each winding element ($\vec{E}_i$) was first obtained as defined in Equations 5.36 and 5.37. Then, the resulting phase EMF phasor ($\vec{E}_{\text{phase,pu}}$) was calculated by summing all the individual winding elements in the same phase (see Equation 5.38). The winding factors for different pole and slot combinations can then be calculated by dividing the magnitude of the resulting phase EMF phasor by the number of winding elements in the same phase as given in Equation 5.39.

\[
\vec{E}_i = e^{j(\gamma_i)} \\
\gamma_i = \frac{2\pi p_i}{Q_s} \\
\vec{E}_{\text{phase,pu}} = \sum_{S/3} \vec{E}_i \\
k_w = \frac{|\vec{E}_{\text{phase,pu}}|}{S/3}
\]

where $\gamma_i$: argument of the phase EMF phasor, $p$: pole pairs and $S$: total number of winding elements.

The analytical equations for calculating the winding factor of double-layer fractional-slot concentrated windings are given in [75,146,147]. It was also shown in [75] that the winding factor can be increased by transforming from a double-layer to a single-layer winding while satisfying geometrical and electrical constraints.

### 5.5.2 Slot and Pole Combination Selection

The required back-iron thickness is inversely proportional to the number of poles. In order to maintain the same back-iron thickness flux density, a thicker back-iron is required when using a lower number of poles. However, this reduces the available slot
depth and electric loading of the machine. On the other hand, a high number of poles requires a high number of teeth and hence increasing magnetic leakage effects. In addition, the higher electrical frequency increases the iron loss. Furthermore, the selection of the number of teeth and poles is also restricted by the construction issues of the maximum allowable outer diameter, slot sizes and the range of commercially available magnets.

Fractional-slot PM motors are becoming more popular due to their low cogging torque and copper loss, and their suitability for fault-tolerant and flux-weakening implementations. There are a number of publications in the literature [72–77, 81–83] on radial-field motors utilising fractional-slot design. These studies primarily cover developing methods to obtain the winding layout for fractional-slot, design criteria, characteristics analysis and performance evaluation. It should be noted that a fractional-slot stator winding has a non-integral number of slots per pole per phase, which is commonly known as “q”. Table 5.4 gives a summary of the main features of various fractional-slot designs for radial-field machine summarised in [77]. This table provides a good starting point in selecting a suitable design in radial machines.

However, there has only been limited research in factional-slot axial-field motors in the literature, which have been mainly in double-sided configurations [70, 148, 149]. In addition, the previous studies mainly utilised SMC material due to its simplicity in producing 3D stator designs especially in axial field stators [14, 15, 19, 20, 150]. Table 5.5 summarises the slot and pole combinations studied in these references.

The stator winding designs including single and double-layer were also studied in the literature in detail [74–76, 78, 109, 147, 151].

Table 5.6 summarises the possible combinations of slots and poles for three-phase concentrated windings as given in [109]. In the table, the winding factor values in parentheses indicate combinations that support extension of the tooth tips to concentrate the airgap flux. Note that, this situation occurs when the q is less than 1/3. In addition, the underlined q combinations in the table indicate the values that allow conversion from double-layer to single-layer windings [78, 151].
Table 5.4. Summary of the main features of various slots per pole per phase (q) combinations for fractional-slot designs with concentrated windings [77].

NOTE:
This table is included on page 122 of the print copy of the thesis held in the University of Adelaide Library.

<table>
<thead>
<tr>
<th>Reference</th>
<th>slot (S) and pole (P) combinations</th>
</tr>
</thead>
<tbody>
<tr>
<td>[70]</td>
<td>12S10P</td>
</tr>
<tr>
<td>[148]</td>
<td>6S4P, 10S4P, 10S6P</td>
</tr>
<tr>
<td>[15]</td>
<td>24S28P</td>
</tr>
<tr>
<td>[14]</td>
<td>18S6P</td>
</tr>
<tr>
<td>[150]</td>
<td>21S20P</td>
</tr>
<tr>
<td>[19,20]</td>
<td>15S14P</td>
</tr>
</tbody>
</table>
As explained in [109], the optimum slot/pole combination for fractional-slot machine can be chosen using the following guidelines, which accommodates a number of tables (Table 5.4, 5.6, 5.7 and 5.8).

- a high value of winding factor, $k_{w1}$ (refer Table 5.6).
- a high LCM (lowest common multiple) of the number of slots and the number of poles (refer Table 5.7).
- a high even value of $K = \frac{(S \times 2p)}{\text{LCM}(S, 2p)} = \text{GCD}(S, 2p)$ where $S$ is number of slots, $p$ is number of pole pairs and GCD is the greatest common divisor (refer Table 5.8).
- choice of single versus double-layer windings depending on application, but only values of $q$ underlined in Table 5.6 are compatible with single-layer windings.
- selection based on features and applications as given in Table 5.4.

As it is known, the fundamental winding factor ($k_{w1}$) determines the number of effective turns in the stator winding coils. Therefore, a high value of $k_{w1}$ leads to better utilisation of the copper winding and hence minimising copper loss.

In addition, the high LCM values of the number of slots and the number of poles would lead to high cogging torque frequency and lower cogging torque amplitude. This is important in concentrated winding designs especially with large slot pitch where skewing may not be effective or practical [109].

The value of $K$ in Table 5.8 shows the symmetry of the machine. A high even value of $K$ means that the machine has symmetrical excitation of the stator coils. Single or double-layer winding configurations each have their own advantages depending on the nature of the application. However as mentioned above, only certain combinations are possible with single-layer designs.
Table 5.6. Number of slots per pole per phase \((q)\) and winding factor \((k_w)\) for various combinations of stator slots \((S)\) and poles \((P)\) with balanced three-phase concentrated windings [109].

NOTE:
This table is included on page 124 of the print copy of the thesis held in the University of Adelaide Library.

Table 5.7. Evaluation of LCM(S,P) parameter for potential slot/pole combinations [109].

NOTE:
This table is included on page 124 of the print copy of the thesis held in the University of Adelaide Library.
5.5.3 Winding Design

After the number of slots and poles have been selected the number of turns and diameter of the copper wires can be found utilising the fundamental induced back-EMF equation.

The number of series-connected turns per phase $N_{Ph}$ can be calculated using the principal equation of induced voltage $E$ which can be defined based on the DC link voltage of the motor drive.

$$E = 4.44N_{Ph}\Phi_{pk}k_{w1}f$$ \hspace{1cm} (5.40)

where $f$: electrical frequency in Equation 5.41 and $\Phi_{pk}$: peak magnetic flux per pole due to magnet from Equation 5.18.

$$f = \frac{np}{60}$$ \hspace{1cm} (5.41)

where $n$: speed in rpm.

After obtaining the number of turns per phase in Equation 5.40, the number of turns per coil $N_c$ can be found using the nearest integer less than or equal to the equations given below.
For a single-layer winding

\[ N_c = \frac{a_p N_{ph}}{p q} \]  

(5.42)

For a double-layer winding

\[ N_c = \frac{a_p N_{ph}}{2 p q} \]  

(5.43)

where \(a_p\): number of parallel current paths and \(q\): number of slots per pole per phase.

The design value of the rms induced voltage \(E\) can be found based on the machine’s rated voltage (\(V_{Lpk}\) peak line-to-line voltage, see Equation 5.44) if given in the design specification. It also can be determined from the DC link voltage (\(V_{DC}\)) of a three-phase voltage-source six-step inverter. In addition, the inverter output voltage waveform can use a 120° or 180° switch conduction pattern. Table 5.9 gives a summary of the relationships between the DC link voltage \(V_{DC}\) and the rms value of the fundamental output voltage for both phase and line voltages [110,152].

### Table 5.9.
Maximum fundamental AC output rms voltage available from a given DC link voltage using 120° or 180° six-step operation [110,152].

<table>
<thead>
<tr>
<th>Conduction Period</th>
<th>Fundamental Voltage (rms)</th>
</tr>
</thead>
</table>
| 120°              | \(\text{Phase Voltage} = \frac{\sqrt{6}}{2\pi} V_{DC}\)  
                   | \(= 0.390 V_{DC}\)  
                   | \(\text{Line Voltage} = \frac{3}{\sqrt{2\pi}} V_{DC}\)  
                   | \(= 0.673 V_{DC}\) |
| 180°              | \(\text{Phase Voltage} = \frac{\sqrt{2}}{\pi} V_{DC}\)  
                   | \(= 0.450 V_{DC}\)  
                   | \(\text{Line Voltage} = \frac{\sqrt{6}}{\pi} V_{DC}\)  
                   | \(= 0.780 V_{DC}\) |

In order to take into account the finite stator winding inductance of the machine, the ratio of the phase EMF to phase voltage ratio should be somewhat less than 1 for motoring mode and somewhat greater than 1 for generating mode operation.

\[ E \approx \frac{V_{Lpk}}{\sqrt{3}\sqrt{2}} \]  

(5.44)
Chapter 5  110mm Machine Analytical Analysis and Design

The size of the wire can be determined based on available stator slot area. The area of a stator slot $A_{\text{slot}}$ was estimated assuming the coil filled an area of rectangular shape of width ‘W’ and height ‘H’ (see Figure 5.16), and the net copper area $A_{\text{cu}}$ was estimated using the packing factor $p_{f,\text{slot}}$ (typical of 0.30-0.35) chosen as given below in Equation 5.45. For the double-layer configuration, half of length ‘W’ was used to calculate the available slot area. The required number of turns was then used to determine the wire area $A_{\text{wire}}$ and wire diameter $D_{\text{wire}}$ as given below in Equations 5.46 and 5.47.

$$A_{\text{cu}} = A_{\text{slot}} p_{f,\text{slot}}$$  \hspace{1cm} (5.45)

$$A_{\text{wire}} = \frac{A_{\text{slot}}}{N_c}$$  \hspace{1cm} (5.46)

$$D_{\text{wire}} = 2\sqrt{\frac{A_{\text{wire}}}{\pi}}$$  \hspace{1cm} (5.47)

![Figure 5.16. Diagram illustrating the slot width and height for a double-layer concentrated winding.](image)

5.6 Resistance Calculation

The DC electrical resistance of the stator winding can be calculated from Equation 5.48. As the resistance of metals increases with temperature, the operating temperature affects the DC resistance of the winding as given in Equation 5.49. In addition, the resistance in practical machines increases with alternating AC current due to the skin and proximity effects which is a function of the skin depth $\delta$ (see Equations 5.50 and 5.51) [153]. Note that, Equation 5.50 can only be applied for rectangular slots with a minimum of 5 winding layers.
\[ R_{wire} = \frac{l_{wire}}{a_p \sigma_{cu} A_{wire}} (N_{cSeries}) \]  
\[ R_{DCwire}(T) = \frac{l_{wire}(1 + k_{cu}(T - 20))}{a_p \sigma_{cu} A_{wire}} (N_{cSeries}) \]  
\[ R_{ACwire}(T) = R_{DCwire}(T) \left[ 1 + \frac{1}{9} \left( \frac{\Gamma_{depth}}{\delta} \right)^2 \left( \frac{D_{wire}}{\delta} \right)^2 \right] \]  
\[ \delta = \frac{1}{\sqrt{\pi} f \mu_0 \sigma_{cu}} \]  

where \( l_{wire} \): length of wire in a coil, \( \sigma_{cu} \): electrical conductivity of copper at 20°C, \( N_{cSeries} \): number of coils connected in series in a phase, \( k_{cu} \): temperature coefficient of the conductivity of copper \( 3.8 \times 10^{-3} \) per °C, \( T \): average winding temperature in °C, \( \delta \): skin depth and \( \Gamma_{depth} \): stator tooth height that is equal to \( H \) in Figure 5.16.

For the design considered, the length of wire per coil \( (l_{wire}) \) equals to \( N_c \times MLT \), where \( MLT \) is the mean length per turn. The \( MLT \) is estimated from the mid point of the coil as shown in Figure 5.17.

**Figure 5.17.** Mean length of the winding per turn.

The winding radial length was approximated from the inner and outer radii of the winding at the mid point of the winding (see Equations 5.52 and 5.53). Similarly, the inner and outer circumferential winding length include the extra length at the mid point radii (see Equations 5.54 and 5.55). The parameter \( a \) in the equations equals to 2 for a double-layer winding and equals to 1 for a single-layer winding. Hence, the mean length of the wire per turn \( MLT \) can be given by Equation 5.56.
\[ Ext = \Gamma_{slot} \frac{2}{a} \]  
\[ WST = 2 \left( (R_{StaOut} + Ext) - (R_{StaIn} - Ext) \right) \]  
\[ WIn = \left( \Theta_{toothIn} + \frac{\Theta_{slotIn}}{a} \right) (R_{StaOut} - Ext) \]  
\[ WOut = \left( \Theta_{toothOut} + \frac{\Theta_{slotOut}}{a} \right) (R_{StaOut} + Ext) \]  
\[ MLT = N_c (WIn + Wout + 2WST) \]  

### 5.7 Phase Inductance Calculation

The synchronous inductances are important parameters to estimate the armature current waveform as they determine the rate of rise or fall of current [110]. The current may not reach the design value as the rate of rise of current is limited by inductance. Consequently, the average current is reduced and hence the rated torque may not be reached. The \(d\)- and \(q\)-axis synchronous inductances are the sum of the armature reaction inductances \(L_{sd}, L_{sq}\) and stator leakage inductances \(L_1\).

\[ L_{sd} = L_{ad} + L_1 \]  
\[ L_{sq} = L_{aq} + L_1 \]  

The calculation of the axis inductances taking into account skewing was presented in [154]. However, the method described in [111] was adapted in this work as the influence of skewing is not considered.

The armature reaction \(d\) and \(q\)-axis inductances are given in given by

\[ L_{ad} = \frac{m \mu_0}{\pi} \left( \frac{N_{Ph}k_{wl}}{p} \right)^2 \frac{R_{StaOut}^2 - R_{StaIn}^2}{g_d'} k_{fd} \]  
\[ g_d' = l_g k_c + \frac{l_m}{\mu_r} \]  
\[ L_{aq} = \frac{m \mu_0}{\pi} \left( \frac{N_{Ph}k_{wl}}{p} \right)^2 \frac{R_{StaOut}^2 - R_{StaIn}^2}{g_q'} k_{fq} \]  
\[ g_q' = l_g k_c + l_m \]
where \( m \): number of phase, \( k_f \): form factors of armature reaction and for the surface magnet configuration of magnets \( k_{fd} = k_{fq} = 1 \).

As given in [111], the stator leakage inductance \((L_1)\) consists of the slot leakage inductance \((L_{1s})\), the end-winding leakage inductance \((L_{1e})\) and the tooth tip inductance \((L_{1tt})\) (see Equation 5.62). The various leakage inductances can be evaluated based on their respective permeance as given in Equations 5.63-5.66.

\[
L_1 = L_{1s} + L_{1e} + L_{1tt} \tag{5.62}
\]

Slot leakage permeance \(\lambda_{1s}\), for a rectangular slot completely filled with conductors:

\[
\lambda_{1s} = \frac{\Gamma_{\text{depth}}}{3 \Gamma_{\text{slot}}} \tag{5.63}
\]

Leakage permeance of the inner \((\lambda_{1\text{ein}})\) and outer end-windings \((\lambda_{1\text{eout}})\), for double-layer windings:

\[
\lambda_{1\text{ein}} \approx 0.17 q \left(1 - \frac{2}{\pi}\right) \frac{w_{\text{cin}}}{l_{1\text{in}}} \tag{5.64}
\]

\[
\lambda_{1\text{eout}} \approx 0.17 q \left(1 - \frac{2}{\pi}\right) \frac{w_{\text{cout}}}{l_{1\text{out}}} \tag{5.65}
\]

where \(w_{\text{cin}}\): inner coil span, \(w_{\text{cout}}\): outer coil span, \(l_{1\text{in}}\): length of inner end connection and \(l_{1\text{out}}\): length of outer end connection.

Tooth tip leakage permeance, \(\lambda_{1tt}\)

\[
\lambda_{1tt} = \frac{5 l_g / \Gamma_{\text{slot}}}{5 + 4 l_g / \Gamma_{\text{slot}}} \tag{5.66}
\]

Hence, the total value of the stator leakage inductance is given by

\[
L_1 = 2 \mu_0 \frac{\Gamma_{\text{StackL}} N_P^2}{Pq} \left( \lambda_{1s} + \frac{l_{1\text{in}}}{\Gamma_{\text{StackL}}} \lambda_{1\text{ein}} + \frac{l_{1\text{out}}}{\Gamma_{\text{StackL}}} \lambda_{1\text{eout}} + \lambda_{1tt} \right) \tag{5.67}
\]

where \(\Gamma_{\text{StackL}}\): stack length \((R_{\text{StaOut}} - R_{\text{StaIn}})\).

The analytical, FE simulated and measured inductances comparisons will be further discussed in Section 7.1.4.
5.8 Phase Voltage and Current

The first approximation of the phase terminal voltage $V$ can be estimated by utilising Equation 5.40, which is also given by

$$V \approx E = 4.44N_{ph}\Phi_{pk}k_{w1}f$$

(5.68)

The maximum phase current ($I_{phMax}$) can be determined from the wire’s current density as given below.

$$I_{phMax} = J_{wire}a_p$$

(5.69)

Typical values of maximum current density are given in Table 5.10 which can be accommodated in the design.

<table>
<thead>
<tr>
<th>Condition</th>
<th>Current Density $J(A/mm^2)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Totally enclosed</td>
<td>1.5-5</td>
</tr>
<tr>
<td>Air-over, fan cooled</td>
<td>5-10</td>
</tr>
<tr>
<td>Liquid cooled</td>
<td>10-30</td>
</tr>
</tbody>
</table>

5.9 Calculation of Losses

5.9.1 Core Loss

The core loss of the stator can be estimated using the core loss formula obtained from the ring core iron loss tests conducted previously. As given in Section 3.4, based on the coefficients extracted from the tests, the formula provides the specific loss ($W/kg$) of the core as a function of peak flux density and frequency. Then, the specific loss results can be multiplied with the respective weights of iron in the teeth and yoke to obtain the loss in watts.

The iron loss depends on the teeth and yoke flux densities in the core and excitation frequency. The frequency is easily obtained from Equation 3.8. As for the flux densities, the peak values can be estimated using the methods given earlier in Sections 5.2 and 5.3. For non-sinusoidal flux density waveforms, the eddy-current loss can be found from the rms value of the rate of change of flux density in the Equation 3.9.
5.9.2 Copper Loss

The copper loss in the stator winding can be approximated using the phase resistance \( R_{\text{wire}} \) calculated in Section 5.6, and using the phase current in Section 5.8.

\[
P_{\text{culoss}} = 3I_{\text{ph}}^2R_{\text{wire}}
\]

Alternatively, the copper loss can also be estimated based on current density \( J \), electrical resistivity of copper \( \rho_{\text{cu}} \) and volume of copper \( \text{Vol} \) as given below. Note that for copper wire, \( \rho_{\text{cu}} = 1.72 \times 10^{-8} \, (\Omega \text{m}) \) at 20°C

\[
P_{\text{culoss}} = J^2 \rho_{\text{cu}} \text{Vol}
\]

Then, the volume of the wire can be estimated by the equation given below. Note that this equation utilises the winding size described in Section 5.5.3.

\[
\text{Vol} = N_{\text{Toothwire}} \times A_{\text{wire}} \times MLT \times N_c
\]

where \( N_{\text{Toothwire}} \): number of teeth wound with wire and \( MLT \) (see Equation 5.56).

If the winding information is not available however, a rough copper loss value can be obtained using the copper volume approximated from Equation 5.73, which neglects the end-windings.

\[
\text{Vol} = Q_s \times A_{\text{slot}} \times p_{f_{\text{slot}}} \times \Gamma_{\text{StackL}}
\]

5.9.3 Magnet and Rotor Core Losses

The eddy-current losses in the rotors of surface-mounted permanent magnet machines are usually neglected. Nevertheless, this assumption is not applicable for fractional-slot designs with surface-mounted permanent magnets equipped with concentrated windings.

The high amplitude spatial subharmonics produced by the stator windings can induce large eddy-current losses in the rotor especially at high speeds. Several analytical models were developed in [85–93, 111] addressing this issue for radial-field machines. A comprehensive literature review on rotor eddy-current losses and the analytical models developed for radial-field machines was also reported in [90].

The main sources of eddy-current losses in the rotor are the space harmonics due to the slotting effects and the stator winding distribution, and the stator current time harmonics. The model reported in [91] is able to predict the eddy-current loss due to
low-order spatial harmonics which is critical for fractional-slot machines. This model was implemented effectively in [155] to investigate the losses for a fractional-slot machine with concentrated windings. In addition, the same model can also be used to calculate the eddy-current loss in the back-iron. In order to take into consideration of the stator current time harmonics, the model was further extended in [89].

The classical formula to estimate the eddy loss in laminated rotor due to the fundamental armature flux was given in [80]. An analytical equation was also derived for the losses in the solid back-iron of concentrated fractional-pitch windings machines in [156,157]. It was shown that the back-iron loss was underestimated at low frequencies and overestimated at high frequencies.

In [112], a simple expression derived from semi-analytical and static finite element solutions was reported. It included the contributions from the stator slot harmonics and the winding space harmonic flux ripple. However, the final equation was limited to the cases when the eddy-current skin depth in the magnet at the relevant harmonic frequency is larger than the magnet thickness. A similar expression was derived for a large horsepower axial field motor in [112,158] which included the effect of skin depth parameter. However, both the current time harmonics and the winding space harmonics were neglected.

As the analytical investigation of eddy-current loss was not the main scope of this work, it was investigated with 3D finite element analysis only in Section 8.3.1.

### 5.9.4 Mechanical Rotational Losses

The mechanical rotational losses considered are the friction loss in the bearing and the windage loss from the rotating rotor disc. According to the reference [111], the rotor windage losses can be estimated using Equation 5.74 for air cooling,

\[
P_{\text{WindLoss}} = 0.5 \ c_f \ \rho_{\text{air}} \ \frac{\omega_M}{3} \ (R_{\text{RotOut}}^5 - R_{\text{shaft}}^5)
\]

Here \( c_f \) is the coefficient of drag for turbulent flow is given by

\[
c_f = \frac{3.87}{\sqrt{Re}}
\]

In the above equation, \( Re \) is the Reynolds number for a rotating disc given by

\[
Re = \frac{\omega_M \ \rho_{\text{air}} \ R_{\text{RotOut}}^2}{\eta_{\text{air}}}
\]
where $\rho_{\text{air}}$: air density at 1 atm and 20°C (1.2 kg/m$^3$), $\eta_{\text{air}}$: dynamic viscosity of air at 1 atm and 20°C (1.8x10$^{-5}$ Pa·s), $\omega_M$: mechanical rotational speed in rad/s and $R_{\text{shaft}}$: radius of the shaft.

The friction loss in a small bearing can be roughly estimated as given below [111].

$$P_{\text{FricLoss}} = 0.03 k_f b \left( \text{Rotor Mass} + \text{Shaft Mass} \right) \frac{\omega_M}{\pi}$$ (5.77)

where $k_f b = 1$ to $3 \text{ m}^2 / \text{s}^2$ for small bearings and the masses are given in kg.

Note that, bearing loss calculation tools from the bearing manufacturer can provide a more accurate loss estimation. In addition, the bearing loss depends on the axial and radial loads on the bearing. The large axial force between the rotor and stator of a single-sided AFPM machine results in a large axial loading on the bearing and hence higher bearing loss. It is possible to lower the axial loading by counterbalancing the force in the double-sided machine configuration. On the other hand, the radial component depends on the weight of the rotor and shaft which can be high for rotors with thick back-iron. Therefore, the bearing should be chosen carefully in AFPM machine design so that they will be able to withstand both the required axial and radial loading with acceptable bearing losses. Further comparison of the different calculations are given later in Section 8.2.

## 5.10 Efficiency

The machine efficiency is approximated from the calculated output power (Equation 5.78) and input power (Equation 5.79) as shown in Equation 5.80. As shown in the equations, the mechanical output power ($P_{\text{Out}}$) is estimated using the calculated torque (see Section 5.4.2) and the mechanical rotational speed ($\omega_M$), and the input power is estimated from the sum of the output power and the losses as described in the previous section.

$$P_{\text{Out}} = \frac{T_M}{\omega_M}$$ (5.78)

$$P_{\text{in}} = P_{\text{Out}} + P_{\text{FEloss}} + P_{\text{cullossR}} + P_{\text{Eddyloss}} + P_{\text{WindLoss}} + P_{\text{FricLoss}}$$ (5.79)

$$\eta = \frac{P_{\text{Out}}}{P_{\text{in}}}$$ (5.80)
Chapter 5  110mm Machine Analytical Analysis and Design

5.11 Design of the 110mm Prototype Machine

The machine design involves determination of the size, slot and pole combination, magnet thickness, slot depth (SD), slot width (SW), stator yoke thickness and winding turns while taking into account of the cutting cost of AMM material and its magnetic properties (i.e. low saturation point). Figure 5.18 shows a flow chart of the general design procedure and analysis process for an AFPM machine based on AMM. The following subsections describe the prototype design procedure according to the flow chart given.

As indicated previously, an AMM-based AFPM motor design is considered in this research. Some of the construction, test setup, material and test machine specifications are listed below.

- Maximum allowable stator outer diameter is 120mm.
- Available three-phase inverter, (340V_{dc} input, 1,000W output power).
- Maximum dynamometer speed of 10,000rpm.
- Neodymium sintered magnets
- Maximum allowable copper current density J (A/m^2, 5x10^6)
- Achievable slot packing factor pf_{slot} (typically 35%)

In the following subsections, these specifications will be discussed to demonstrate the approach taken towards the final machine prototype.

5.11.1 AMM Material Considerations

It is important to note here that there are two critical issues when AMM is considered in an AFPM machine design: width of available AMM ribbon and slot cutting rate.

Since AMM is produced in ribbon form, there are practical manufacturing limitations about the maximum width of the raw AMM ribbon. It was provided by Metglas at the time of design that the maximum width was 212mm. This research study utilised 30mm AMM ribbon in the design as it suited to the targeted motor size.

Due to the cutting technique utilised by the industrial partner, there is a trade-off between the performance and cutting cost. The cutting rate slows down substantially
Chapter 5  110mm Machine Analytical Analysis and Design

Determine the outer and inner diameter of the machine based on application, sizing equations, size of available AMM ribbon and AMM cutting constraints.

Select the optimum slot/pole combination.

Determine the air gap length, magnet type and thickness.

Determine the slot depth and thickness of stator back iron (saturation point affecting torque for tooth and yoke flux density based on AMM).

Determine single or double layer winding design; calculate wire size and number of turns per phase and per coil.

Calculate machine’s parameters: back-EMF constant, resistance, inductance.

Calculate machine’s performance: torque, efficiency, main losses: iron, copper, eddy-current in rotor, bearing and windage.

Does the machine meet the specifications?

Yes

Compare with 3D FEA for final validation.

No

Adjust values.

Prototype construction and testing.

Figure 5.18. Analytical design flow chart.
with increasing cutting thickness (difference between inner and outer radius). In addition, the length of the cutting path increases faster with slot depth compared to the slot width. As a result, a deeper slot depth would increase the cutting time significantly compared to a wider slot width.

In general, the lower number of slots would require less cutting time and hence reduces the cost. Nevertheless, a machine with better performance would generally require a high pole number which leads to higher number of slots.

5.11.2 Sizing

As stated before, the axial-field machine size was determined by the available toroidal shape non-slotted core samples provided by the industrial partner for the laboratory prototype. The samples have 110mm outer diameter and 45mm inner diameter which results in a 0.41 diameter ratio. It should be noted that the diameter ratio is not optimised and will be considered in future work.

Figure 5.19 illustrates three uncut non-slotted cores: SMC, SI and AMM. The SI and AMM cores were wound with ribbon material of 30mm width. The available raw SMC core had the same thickness as the SI and AMM cores but had smaller inner diameter.

The inner diameter of 45mm was found sufficient to accommodate a reasonable diameter shaft which can support the magnets and rotor core at high speed operation and to accommodate winding ends. The inner section of the raw SMC was machined to
obtain larger inner diameter. Then, all the cores were machined or cut to make suitable number of slots.

### 5.11.3 Slot and Pole Combination

A previously constructed SMC-based AFPM prototype machine was used during this study. The stator core had been made by precision machining from a cylindrical piece of SMC (see Figure 5.19(a)) of outer diameter $110\, mm$, inner diameter $30\, mm$ and has a stator back-iron of thickness $10\, mm$. This prototype has twelve teeth, each wound with a concentrated winding using square cross-section copper wire. The four pole rotor consists of four arc-shaped, surface-mounted magnet segments with a solid mild steel back-iron. It can be noted that the choice of four poles does not appear to be optimum as such low pole number is likely to cause saturation in the stator and rotor back-iron. The prototype machine uses an unusually large airgap, $6.5\, mm$ compared with the $2.5\, mm$ magnet thickness. Possible reasons for this may include: to reduce magnetic flux densities in the stator to avoid saturation or reduce stator iron losses; or to reduce the axial force on the machine to reduce bearing losses.

As the raw AMM toroidal core has a similar outer dimensions as the SMC machine, the AMM stator was initially designed to use the above 4 pole PM rotor. With the design airgap length of $1\, mm$, the AMM stator and rotor back-iron thicknesses were increased to avoid excessive saturation.

It should be noted that it became possible at later stage to order various custom designed magnets and hence other slot and pole designs were explored. A variety of slot and pole combinations and their resulting magnetic loading were investigated in [74]. Among those, the $12S10P$ is one of the few combinations that can support single or double-layer windings, which also offers high winding factor, low torque ripple and no mutual coupling. As mentioned previously, the higher pole number allows a smaller yoke thickness. In addition, the high inductance of such fractional-slot design is suitable for wide constant power speed range operation and limits the short-circuit fault current. The absence of mutual coupling makes the $12S10P$ combination suitable for fault tolerant applications.

Fractional-slot machines with $q$ between 0.33 and 0.5 ($12S10P$ has $q=\frac{12}{10\times3} = 0.4$) generally have high performance according to [77]. This was investigated and verified
experimentally in [78, 147]. Also, these configurations can achieve a high torque density as demonstrated in [147]. Therefore, the 12S10P combination was chosen as the reference model, and compared to the 12S4P combination with identical stator total depth and magnet thickness.

Based on the reference base model (see Table 5.1), the output torque (Equation 5.24), back-EMF (Equation 5.68) and iron loss (Equation 3.8) were calculated analytically for comparison. Table 5.11 shows the designed slot depth and width based on maintaining the tooth and yoke flux densities within the 1.57\(T\) saturation flux density of the AMM material. In addition, Table 5.12 is given to summarise the calculated back-EMF, torque and iron loss. As can be seen from the table, the back-EMF is 70\% higher and the torque of the 12S10P design is about 2 times larger compared to the 12S4P design. However, the iron loss is also 3 times higher, and it should be noted that the increased slot depth in the 12S10P design would increase the cutting cost substantially.

**Table 5.11.** Summary of analytically calculated tooth (Model B2) and yoke flux densities (Model By2) for 12S10P and 12S4P combinations.

<table>
<thead>
<tr>
<th>Combinations</th>
<th>Slot Depth (mm)</th>
<th>Slot Width (mm)</th>
<th>Tooth flux density (T)</th>
<th>Yoke flux density (T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>12S10P</td>
<td>24</td>
<td>8</td>
<td>1.36</td>
<td>1.39</td>
</tr>
<tr>
<td>12S4P</td>
<td>12</td>
<td>8</td>
<td>1.36</td>
<td>1.30</td>
</tr>
</tbody>
</table>

**Table 5.12.** Summary of analytically calculated torque, back-EMF and iron loss for 12S10P and 12S4P combinations at 10,000rpm.

<table>
<thead>
<tr>
<th>Combinations</th>
<th>Back-EMF (V)</th>
<th>Peak Torque (Nm)</th>
<th>Iron Loss (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>12S10P</td>
<td>185</td>
<td>5.40</td>
<td>32</td>
</tr>
<tr>
<td>12S4P</td>
<td>108</td>
<td>2.70</td>
<td>11</td>
</tr>
</tbody>
</table>

It can be concluded that from the peak torque equation (see Equation 5.23), a linear relation with the number of slots and the slot size can be deduced. The peak torque increases as thicker wire capable of handling higher current can be packed into the stator with larger slot size. The number of poles has no direct effect on the peak torque.
Chapter 5 110mm Machine Analytical Analysis and Design

and copper loss. However it does affect the back-iron thickness and hence allowable slot depth.

In addition, since iron loss is proportional to both frequency and flux density squared, it is affected by the pole number which changes the yoke flux density and the operating frequency, and also by the slot width which affects the tooth flux density. Further experimental comparisons will be given later in Section 7.4.2.

### 5.11.4 Slot Depth and Slot Width

With regards to the magnetic loading, the slot depth and width are determined by the maximum allowable stator tooth and yoke flux density which is about 1.57T for AMM. Based on the analytical calculation methods described in Sections 5.2 and 5.3, the estimated stator dimensions are given in Table 5.13. As described in Section 5.4.1, since the tooth flux density varies with radial position, the values at the average and outer radii were given in the table. As the saturation point of AMM is lower compared to conventional iron, the tooth flux density for similar size can exceed the material’s saturation point. Even so, the machine could still maintain its output performance with slight saturation. This will be investigated further in Section 6.4.2 using 3D FE analysis.

As it is known, with regards to electric loading, there should be a reasonable space for copper. The double-layer winding arrangement was chosen in the design. A slot width of 8\(\text{mm}\) was found to keep the tooth flux density within the saturation value of AMM at the average radius. In addition, it was found that a slot depth of 24\(\text{mm}\) and a base height of 6\(\text{mm}\) would satisfy the magnetic loading for the yoke. However the limitation of the cutting facility did not allow the slot depth to be greater than 17\(\text{mm}\).

Hence, the 17\(\text{mm}\) slot depth AMM stator core design was chosen. However, the 17\(\text{mm}\) slot depth stator was not available for testing. Thus, the 12\(\text{mm}\) slot depth stator designed for the 4 pole machine was also used in the 10 pole machine. The small number of poles required a thicker stator yoke to avoid excessive saturation. Therefore, the 12\(\text{mm}\) slot depth was used in the reference model as given in Table 5.1. In Section 7.1.1, the performance of 12\(\text{mm}\) and 17\(\text{mm}\) slot depth stators will be compared utilising 3D FE analysis. It should be noted that all the test results in Chapters 7 and 8 were based on the 12\(\text{mm}\) slot depth stator.
Table 5.13. Analytically calculated yoke and tooth flux densities (B) at various slot widths, slot depths and yoke heights.

<table>
<thead>
<tr>
<th>Slot Depth (mm)</th>
<th>Base Height (mm)</th>
<th>Slot Width (mm)</th>
<th>B stator yoke (T)</th>
<th>B tooth at average radius (T)</th>
<th>B tooth at outer radius (T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>24</td>
<td>6</td>
<td>4</td>
<td>1.52</td>
<td>1.13</td>
<td>1.06</td>
</tr>
<tr>
<td>24</td>
<td>6</td>
<td>6</td>
<td>1.46</td>
<td>1.24</td>
<td>1.13</td>
</tr>
<tr>
<td>24</td>
<td>6</td>
<td>8</td>
<td>1.39</td>
<td>1.36</td>
<td>1.18</td>
</tr>
<tr>
<td>24</td>
<td>6</td>
<td>10</td>
<td>1.30</td>
<td>1.54</td>
<td>1.27</td>
</tr>
<tr>
<td>17</td>
<td>13</td>
<td>4</td>
<td>0.70</td>
<td>1.13</td>
<td>1.06</td>
</tr>
<tr>
<td>17</td>
<td>13</td>
<td>6</td>
<td>0.68</td>
<td>1.24</td>
<td>1.13</td>
</tr>
<tr>
<td>17</td>
<td>13</td>
<td>8</td>
<td>0.64</td>
<td>1.36</td>
<td>1.18</td>
</tr>
<tr>
<td>17</td>
<td>13</td>
<td>10</td>
<td>0.60</td>
<td>1.54</td>
<td>1.27</td>
</tr>
<tr>
<td>12</td>
<td>18</td>
<td>4</td>
<td>0.51</td>
<td>1.13</td>
<td>1.06</td>
</tr>
<tr>
<td>12</td>
<td>18</td>
<td>6</td>
<td>0.49</td>
<td>1.24</td>
<td>1.13</td>
</tr>
<tr>
<td>12</td>
<td>18</td>
<td>8</td>
<td>0.46</td>
<td>1.36</td>
<td>1.18</td>
</tr>
<tr>
<td>12</td>
<td>18</td>
<td>10</td>
<td>0.43</td>
<td>1.54</td>
<td>1.27</td>
</tr>
</tbody>
</table>

Effect of number of slots on cutting length

Although it is not a common combination for AFPM machines (see Table 5.5), the 12S10P combination was chosen. The 12 slots is a good starting point as it is the mid range of the combinations in table. This is because the total AMM cutting length and time increases linearly with the number of slots, which results in a higher cutting cost.

Figure 5.20 shows the analytical results of the total cutting length as a function of slot numbers at various slot depth and slot width combinations. Note that, SD12 means a slot depth of 12mm and SW8 means a slot width of 8mm. Also note that, for the 12S10P configuration with 45mm inner and 110mm outer diameter, the small inner diameter allows only limited space for slot width variation (4-10mm) while the slot depth was varied from 6 to 25mm. In addition, based on the cutting route, the total cutting length per slot is more sensitive to slot depth than slot width. Hence, there is only small changes in total cutting length with slot width variations as shown in Figure 5.20(a). On the other hand, the effect of slot depth variations is more influential (see Figure
Therefore, it can be concluded that slot depth variation would have a larger effect in terms of the cutting cost.

![Graph (a) For various slot widths (mm)](image1)

![Graph (b) For various slot depths (mm)](image2)

**Figure 5.20.** Calculated cutting length variation versus number of slots for different combinations of slot width (SW) and depth (SD).

### 5.11.5 Airgap and Magnet Thickness

For the prototype machine design, an airgap length of 0.66\( \text{mm} \) was calculated from Equation 5.34 and it was decided to have an airgap length of 1\( \text{mm} \) to take into account of the high axial force in the single-sided configuration. In addition, such value was found reasonable compared with the designs reported in Table 2.7. As for the magnet thickness, a 2.1\( \text{mm} \) magnet thickness was calculated from Equation 5.35 which is considerably smaller compared to the values of 3\( \text{mm} \) in [15], 4\( \text{mm} \) in [14] and 5.7\( \text{mm} \) in [94]. It was decided to choose an airgap flux density within 0.8-0.9\( T \). Hence, a magnet thickness of 3\( \text{mm} \) was chosen.

For the 10 pole rotor, 10 arc-shaped permanent magnets of identical inner and outer diameters were custom made. The magnets were glued on a solid back-iron made from a ring of mild steel with an 9\( \text{mm} \) thickness for the 10P design and a 16\( \text{mm} \) thickness for the 4P design (based on its saturation flux density). Sintered neodymium magnets were first utilised for its high magnetic properties but it was found in Section 7.4.3 that the magnet eddy-current loss was high and so bonded magnets were later ordered for comparison purposes. The test results and analysis will be presented later in Section 8.3.1.
The resistivity of sintered magnets is around $1.5 \times 10^{-6} \Omega m$ and for bonded magnets, it is about ten times larger, $2 \times 10^{-5} \Omega m$. As the remanent flux density of bonded magnets is smaller compared to sintered magnets, the machine design would have to be modified to compensate for the lower remanent flux density value. Furthermore, the magnet mass is larger due to the increase in magnet thickness. In order to maintain the same rated torque, a larger machine or a higher value of stator current density is generally required. Due to these factors the specific torque density of the bonded magnet design would be lower compared to a similar machine with sintered magnets. As a result, there are tradeoffs associated with using the lower-strength bonded magnets.

5.11.6 Matlab Drawing and Calculation

The analytical calculation approach discussed in the previous sections was implemented in Matlab. The aim of this approach is to rapidly predict the characteristics of an AFPM design, which included a 2D sketch of the machine design showing the top and the side views of the stator and the rotor as shown in Figure 5.21.

The estimated machine parameters in the analytical Matlab tool are given in Table 5.14 for the AMM core that is based on the reference model given earlier in Table 5.1. Note that, the values were calculated neglecting saturation for both sintered and bonded magnet designs. The number of turns were kept the same in both designs. As it is discussed previously, bonded magnets were considered to reduce the magnet eddy-current loss. In addition, since the bonded magnets had a lower remanent flux density, thicker magnets were used.

As it can be seen from the table, the bonded magnet design had lower values of peak flux density, back-EMF, output torque and inductances. As expected, the lower flux density resulted in half the iron loss compared to the sintered magnets design.
Figure 5.21. The top and the side view sketches of a design in the analytical approach using Matlab. Design specifications: 12 slot, 10 pole, 8mm slot width, 12mm and 17mm slot depths.
Table 5.14. Summary of analytical design for the 12S10P AFPM AMM machine: slot depth of 12\text{mm} and slot width of 18\text{mm} and rated speed of 10,000\text{rpm}.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Sintered Magnet</th>
<th>Bonded Magnet</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnet remanent $B_r(T)$</td>
<td>1.14</td>
<td>0.70</td>
</tr>
<tr>
<td>AMM saturation $B_{sat}(T)$</td>
<td>1.57</td>
<td>1.57</td>
</tr>
<tr>
<td>Peak airgap $B_g(T)$</td>
<td>0.93</td>
<td>0.65</td>
</tr>
<tr>
<td>Peak effective airgap $B_{ficg}(T)$</td>
<td>0.83</td>
<td>0.59</td>
</tr>
<tr>
<td>Peak tooth $B_{Tave}$ (average)(T)</td>
<td>1.36</td>
<td>0.98</td>
</tr>
<tr>
<td>Peak tooth $B_{Tout}$ (outer)(T)</td>
<td>1.18</td>
<td>0.84</td>
</tr>
<tr>
<td>Rotor magnet pole flux (mWb)</td>
<td>0.54</td>
<td>0.39</td>
</tr>
<tr>
<td>Stator back-iron $B_{SBI}(T)$</td>
<td>0.46</td>
<td>0.33</td>
</tr>
<tr>
<td>Rotor back-iron $B_{RBI}(T)$</td>
<td>0.76</td>
<td>0.54</td>
</tr>
<tr>
<td>Back-EMF voltage $E_{rms}(V)$</td>
<td>185</td>
<td>133</td>
</tr>
<tr>
<td>Maximum phase current $I_{rms}(A)$</td>
<td>3.2</td>
<td>3.2</td>
</tr>
<tr>
<td>Number of phase turns</td>
<td>96</td>
<td>96</td>
</tr>
<tr>
<td>Number of turns per coil</td>
<td>24</td>
<td>24</td>
</tr>
<tr>
<td>Wire diameter (mm)</td>
<td>0.90</td>
<td>0.90</td>
</tr>
<tr>
<td>Copper skin depth (mm)</td>
<td>2.3</td>
<td>2.3</td>
</tr>
<tr>
<td>Phase resistance $R$ (20^\circ C) $\Omega$</td>
<td>0.27</td>
<td>0.27</td>
</tr>
<tr>
<td>Phase resistance $R_{dc}$ (50^\circ C) $\Omega$</td>
<td>0.32</td>
<td>0.32</td>
</tr>
<tr>
<td>Phase $R_{ac}$ (50^\circ C, skin effect) $\Omega$</td>
<td>0.48</td>
<td>0.48</td>
</tr>
<tr>
<td>Leakage inductance (mH)</td>
<td>0.23</td>
<td>0.27</td>
</tr>
<tr>
<td>$d$-axis armature inductance $L_{ad}(mH)$</td>
<td>0.26</td>
<td>0.15</td>
</tr>
<tr>
<td>$q$-axis armature inductance $L_{aq}(mH)$</td>
<td>0.25</td>
<td>0.18</td>
</tr>
<tr>
<td>Peak torque ($Nm$)</td>
<td>2.68</td>
<td>1.92</td>
</tr>
<tr>
<td>Output power ($kW$)</td>
<td>2.8</td>
<td>2.0</td>
</tr>
<tr>
<td>Iron loss ($W$)</td>
<td>14.0</td>
<td>7.02</td>
</tr>
<tr>
<td>Copper loss ($3I^2R_{dc}$) ($W$)</td>
<td>8.3</td>
<td>8.3</td>
</tr>
<tr>
<td>Copper loss ($3I^2R_{ac}$) ($W$)</td>
<td>14.7</td>
<td>14.7</td>
</tr>
<tr>
<td>Rotation loss ($P_{WindLoss} + P_{FricLoss}$) ($W$)</td>
<td>97.5</td>
<td>120.6</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>95</td>
<td>93</td>
</tr>
</tbody>
</table>
5.12 Demagnetisation

As it is known, when the windings in a stator core are energised, the stator armature reaction field shifts the magnet load line which determines the magnet’s operating point. The operating point is the interception of the load line and the magnet’s demagnetisation curve in the second quadrant (see Figure 5.22). As the armature reaction field changes, the operating point moves along the demagnetisation curve. The knee flux density value $B_k$ is the point at which the B-H curve changes from its linear behaviour and is provided in the manufacturer’s technical data sheet.

If the operating point of a magnet is forced below the knee, the magnet will not be able to recover its original remanent flux density and hence has been demagnetised. Parts of the magnet may be demagnetised if exposed to an excessive stator armature reaction field. Therefore, the aim during the design process is to keep the flux density in magnets above the knee value $B_k$ under worst-case operating conditions [110].

The remanent flux density $B_r$ of the magnet drops and the knee flux density $B_k$ values rises as the temperature increases as shown in Figure 5.22. The magnet is thus more prone to demagnetisation from the stator armature reaction at higher temperatures. Therefore, the calculation of the magnet’s operating point should be based on the BH-curve at the worst-case expected magnet operating temperature which is usually around 60-100°C.

![Figure 5.22. Manufacturer’s demagnetisation curve for sintered neodymium magnet.](image)

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Figure 5.22 also gives the BH-curves of a sintered neodymium magnet for different temperatures including the intrinsic magnetisation curve. In the figure, the load line with zero stator current (line passes through the origin) is plotted with a slope equal to the permeance coefficient (PC) which can be calculated from Equation 5.81.

\[ PC = \frac{1}{f_{LKG} \times \frac{l_m}{l_g} \times \frac{A_g}{A_m}} \quad (5.81) \]

where \( f_{LKG} \): is the leakage coefficient with a typical value of 0.8-0.9, \( l_m \): is the magnet thickness and \( l_g' \): is the effective airgap (including Carter’s coefficient).

Note that, the airgap surface area \( A_g \) is nearly equal to the magnet surface area \( A_m \) for surface magnet rotors. Hence, the approximate PC is given by

\[ PC \simeq \frac{1}{f_{LKG} \times \frac{l_m}{l_g}} \quad (5.82) \]

As illustrated in Figure 5.22, the stator armature reaction shifts the load line to the left, which depends on the level of the peak airgap magnetomotive force (MMF). The peak value of the airgap MMF \( H_{pk} \) is calculated with Equation 5.83 for the machine design based on the rated peak phase current (\( I \)) 4.5\( A_{pk} \) (3.2\( A_{rms} \)) and the number of turns per coil (\( N_c \)) of 24 (see Table 5.14).

\[ H_{pk} = \frac{N_c I}{l_m + l_g'} \quad (5.83) \]

As can be seen in Figure 5.22, the magnet operating point at the rated current value is above the knee point (up to an operating temperature of about 140\(^\circ\)C) which is considered acceptable. In order to check the demagnetisation, the peak airgap MMF is calculated from Equation 5.83 for the highest possible loading current. The peak current values for the magnet operating at the knee point and for different temperatures are given in Table 5.15.

Figure 5.23 shows the axial flux density FE plots as a function of circumferential position at 140\(^\circ\)C for the open-circuit condition and for the case of the maximum allowable current of 9\( A_{pk} \) and for a higher current of 18\( A_{pk} \). As it can be seen in the figure, the airgap flux density is approximately 0.74\( T \) for the open-circuit case, which is consistent with the calculation in Figure 5.22. As the current is in the negative \( d \)-axis, the airgap magnet flux is reduced. It can also be seen in Figure 5.22 that the peak current of 9\( A \) would push the magnet operating point down to the knee point at 140\(^\circ\)C. As shown in the Figure 5.23, when the windings energised the flux density is pushed down to about
Table 5.15. Analytically calculated peak current at demagnetisation knee point for various magnet operating temperatures.

<table>
<thead>
<tr>
<th>Temperature (°C)</th>
<th>Peak Current ($A_{pk}$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>20</td>
<td>132</td>
</tr>
<tr>
<td>60</td>
<td>100</td>
</tr>
<tr>
<td>100</td>
<td>36</td>
</tr>
<tr>
<td>140</td>
<td>9</td>
</tr>
</tbody>
</table>

0.68T according to the 140°C curve which is consistent with Figure 5.22. Note that, at a peak current of 18A, the flux density is reduced to 0.47T where demagnetisation would occur if operated at 140°C.

Figure 5.23. FE calculated open-circuit and maximum allowed current operating conditions at 140°C.

It should be noted that the worst-case magnet operating point in Figure 5.23 occurs over the slot openings. This is because of the open slot configuration of the prototype AFPM machine. This effect will not be explored further in this thesis and is left for future work.
5.13 Thermal Analysis

As discussed in the previous section, the remanent flux density of the permanent magnet material depends on the temperature. Therefore, the amplitude of the induced back-EMF changes with temperature and thereby would affect the performance characteristics of the machine. In addition, the winding insulation and stator copper loss are also temperature dependent. Furthermore, higher temperature may also cause the magnets to disintegrate if the temperature of the glue joint exceeds the critical temperature. Thus, estimation of the operating temperature would improve the accuracy of the performance analysis of the electrical machine.

Thermal analysis of axial-field machines were reported in [111, 112, 144, 159]. Accurate thermal modelling of axial-field machines demands substantial work and it is beyond the scope of this research.
5.14 Conclusions

One of the objectives of this research was to develop an analytical design procedure for designing AFPM machines utilising cut AMM. A set of requirements was developed by taking into consideration the available laboratory test rig, magnets and AMM toroidal sample non-slotted core. Then, a design procedure including initial screening and design guidelines leading to a preliminary design was proposed. It includes determination of the size, slot and pole combination, magnet thickness, slot depth, slot width, stator yoke thickness and winding turns of the machine taking into consideration of the limitations of the AMM cutting techniques and its saturation value. The analytical methods were implemented in Matlab. In addition, a magnet demagnetisation of magnet has been studied.

In this chapter, the analytical modelling of surface-mounted AFPM machines with concentrated windings was presented. The aim of the analytical approach is to determine the basic parameters (i.e. number of turns, resistance, inductance) and to predict performance characteristics (i.e. flux density, back-EMF, torque, rotational, iron and copper losses) using the fundamental electromagnetic equations for preliminary design.

The analytical methods chosen were presented and compared with finite element results. It was shown that analytical approximations do not offer accurate design data. However, the analytical approximation approach provides a faster initial design which could be fine tuned and optimised in the final design process. In addition, this work focuses on 3D FE modelling which can achieve high accuracy analysis as will be described in Chapter 6. Although this requires substantial work effort in modelling and simulation, it is the most appropriate tool to analyse and verify the design procedure as far as this research is concerned.

In the chapter, the optimum values of inner/outer diameter ratio, slot depth and width were investigated based on shear stress and iron loss analysis. This was conducted for sizing purpose and to investigate possible options for reducing the cutting cost in terms of smaller stack length (thickness) and slot size (total cutting length). Then, the stator outer diameter was used to estimate initial values of the length of the airgap and the thickness of neodymium magnet which affect the overall axial length of the machine.

The fractional-slot concentrated winding AMM-based AFPM machine was chosen, and guidelines were provided to help the design process. The 12 slot 10 pole design
was chosen in this study as it offers high winding factor and low torque ripple. In addition, it is compatible with single or double-layer windings, has no mutual coupling and has small yoke thickness. However, it was shown that the cost and performance of a design is strongly affected by the number of slots and poles. Increasing the number of slots increases the electric loading and hence torque, and increasing the number of poles decreases the required yoke thickness but also increases the iron loss. In addition, it is concluded that the AMM cutting cost increases with the radial cutting thickness and the cutting length. The cutting length is also more sensitive to the slot depth compared to the slot width. Therefore, it can be concluded that there is a trade-off between the performance and cost with the optimum value depending on the requirements and application.
FINITE-element Analysis (FEA) is able to accurately analyse complex electromagnetic field models based on Maxwell equations. The partial differential and integral equations in Maxwell can be accurately solved in computer programs. Hence, FEA is commonly used to model the magnetic behavior of an electrical machine accurately and predict the flux distribution within the machine in details especially in 3D. In this chapter, modelling of the AFPM machine in 3D based on the JMAG-studio software package is described in detail. The simulated machine’s parameter and performance characteristics are also presented. In addition, the simulated results are compared with experimental results to examine the accuracy of the FE model developed.
6.1 Introduction

There have been a number of publications utilising 2D FEA to both validate analytical models and also to optimise the performance of radial-field machines. However, 3D FEA is required for axial-field PM machines due to the inherent 3D geometry and flux distribution. In addition, 3D analysis offers the best accuracy as all possible flux paths and distributions are taken into consideration. However, it requires lengthy simulation times to generate a fine mesh and to compute the field for specifically the complex 3D structures. Therefore, this has motivated the development of modelling AFPM machine based on analytical and 2D FEA. Although, these methods generally have low accuracy and are usually limited for simple magnet shape cases, they are still acceptable at the initial stage of design especially with tight time constraints as reported in [160]. Nevertheless, it can be concluded that the highly accurate prediction in 3D FEA is suitable for analysing, performance prediction and theoretical concept verification specifically in AFPM machine design. Some recent research work on AFPM machine analysis utilising 3D FEA are reported in [161–167].

In general, a FEA software package consists of three main components which are the pre-processor, field solver and post-processor. The pre-processor involves the creation of the model geometry, material definitions, boundary condition assignments and meshing. After the assignments, the software computes the field and perform electromagnetic analysis on each element of the mesh. The analysis results such as magnetic field, force, torque, flux and inductance etc. are presented in the post-processor in various display forms. The results also provide detailed information on the flux distribution in the machine which can be displayed in contour or vector views. Therefore, the parameter and performance characteristics of the designed machines can be predicted.

In this work, “JMAG-studio” (JMAG) package which supports both 2D/3D FEA by JRI Solutions, Ltd was employed as the design and validation tool. In addition, JMAG was utilised in a PC with AMD Athlon 64 X2 Dual Core Processor 4200+ and 2GB RAM. More details of JMAG package can be found in [168–173].
6.2 3D Modelling

The symmetry and periodicity of the physical structure and the flux distribution of an axial-field PM machine can allow for possible quarter and half modeling in JMAG instead of a full model (see Figure 6.1). Less computation time is needed to perform meshing and analysis for smaller models. Hence, a quarter model is preferable especially in time consuming 3D FEA.

The symmetry and periodicity of the model depends on the configuration of slots (or teeth) and poles of the machine. For example, a full model is needed for a 3 slot 2 pole (3S2P) machine (Figure 6.1(c)), while a quarter model can be used for a 12 slot 4 pole (12S4P) machine (periodicity of 90 degrees, see Figure 6.1(a)) and a half model is the smallest for a 12 slot 10 pole (12S10P) machine (Figure 6.1(b)). Note that, in the figure the rotor back-iron in all models and one of the magnet poles in the half model in Figure 6.1(b) were hidden for viewing purposes.

![Figure 6.1. FE modelling: (a) quarter model, (b) half model, (c) full model.](image)

Due to the 3D geometry of axial-field PM machines and the lack of advanced CAD functions in the JMAG geometry editor, it would take a long time to produce the geometry using the built-in editor. JMAG has a built-in link to SolidWorks and supports various import geometry model formats. In this work, the 3D model was first generated using the Solid Edge package and the model geometry in ‘SAT’ format was imported into JMAG. Figure 6.1 shows the solid models created in Solid Edge, and Figures 6.2(a) and 6.3(a) show the models in JMAG environment.

It should be noted that the windings on the teeth in the 3D model require modification in order to set the direction of current flow in the windings. For example, in the concentrated winding configuration, the winding on each tooth is formed from two parts (see Figure 6.2(a)) so that the “in flow” current direction can be set during creation of...
"FEM coil region condition" in JMAG. In addition, the BH-curves and iron loss characteristics for the stator materials were defined based on the iron loss measurements conducted in Section 3.1 and imported into the material database. The JMAG results were obtained from a time-stepping, coupled-circuit 3D FE simulation.

![Solid model view](image1.png) ![Mesh view](image2.png)

Figure 6.2. 3D FEA quarter model of the 12S4P AFPM machine configuration in JMAG: a) solid model view and b) mesh view (8,711 nodes, 34,834 elements).

### 6.2.1 Meshing and Accuracy

In JMAG, the automated meshing of the axial configuration generates a mesh corresponding to the position of the movable part at each analysis step [173]. The meshing of the stator and rotor parts are created once while the airgap between the stator and rotor is re-meshed for each rotor position (patch mesh method). Hence, the flux distribution in the airgap is not stored in the solution as the mesh is constantly changing.

As stated before, the level of meshing size used in the machine model determines the computation time and the accuracy of the simulation results. In general, the finer the mesh size the more accurate the numerical results and smoother the simulated waveforms. However, the computation time will be longer for each step. In order to reduce the computation time, different size meshing can be set on different parts of the model. As a rule of thumb, small meshing size should be used where the magnetic flux concentrates, in its nearby space where the magnetic flux leaks and also where the size and direction of magnetic flux changes considerably [172].

JMAG offers automatic mesh generation tools which includes auto creation and meshing of a surrounding air volume which is normally not constructed in the 3D CAD
model imported. If required, the model can be further fine tuned later to determine the mesh size based on the computation time and the required accuracy of the simulation results.

Figure 6.2(b) shows the mesh view of a quarter model of the 12S4P machine. JMAG’s automated patch mesh method generated a 3mm mesh size for all parts. The calculated back-EMF and flux density waveforms were smooth and sinusoidal. In addition, the waveforms and predicted parameters showed good accuracy with the measured results from a prototype AMM AFPM machine (see Section 6.3). Therefore, finer meshing is found not necessary in this case.

On the other hand, the 12S10P machine requires a half model due to its fractional-slot structure as shown in Figure 6.3. JMAG’s automated patch mesh method generated 5mm mesh size for all parts, Figure 6.3(b).

As can be seen in Figure 6.4, the generated back-EMF waveform has some ripple although this is not observed in the flux waveforms. This indicates some degree of coarseness in the model which is confirmed by the noisy cogging torque and axial force waveforms generated at this mesh size as shown in Figure 6.5. The waveforms were expected to be more sinusoidal, and for the cogging torque, expected to be symmetrical about zero as it will be shown later in Figures 6.16(a) and 6.17(a).
To improve the above waveforms, the convergence of the iterations is checked first. A possible solution is to change the maximum iteration settings such as increasing the number of the non-linear iterations. Note that, this might also increase the simulation time. If there is no improvement, the mesh level needs to be examined. Then, an optimised mesh size producing acceptable accuracy results with the shortest computing time is investigated. Figure 6.6 gives a summary of the suggested steps for obtaining the optimised mesh size for the AFPM model.
Is the iteration converging?  
If not, change default maximum iteration settings as suggested in error message generated.  
Re-run simulation

Yes

Are the waveforms (e.g. back-EMF, magnetic flux, cogging and axial force) generated with “Automatic Meshing” smooth and acceptable?  

No

Set smaller mesh size for the faces of teeth and magnets at the air gap only, re-run simulation.

Is the iteration converging?  
Are the waveforms generated smooth and acceptable?  

Yes

Is simulation time acceptable?  

No

Delete the face mesh size setting and set mesh size for only the stator core similar to the face size found above, re-run simulation.

Is the iteration converging?  
Are the waveforms generated smooth and acceptable?  
Is simulation time acceptable?  

No

Set smaller mesh size for the stator core only, re-run simulation.

Yes

Compare with experimental for final validation

Figure 6.6. Mesh optimisation flow chart.
The mesh resolution of the stator and magnet surfaces facing the airgap significantly influences the cogging torque and axial force waveforms. On the other hand, the induced voltage and magnetic flux waveforms were not affected greatly. It is necessary to obtain an accurate model of the magnetic flux distribution within the airgap area as this is critical in predicting the machine performance [172]. Therefore, setting a finer mesh size on surfaces facing the airgap would result in more accurate results and smoother waveforms. This method is referred as "Method 1". Nevertheless, setting significantly different size of meshing on different parts of the model could also cause the iteration not to converge, especially in the loaded condition, and in some cases might even increase the simulation time.

Alternatively, it is possible to optimise the model by setting a finer mesh size for the stator only and using the automatic meshing feature to obtain an optimised mesh for the remaining parts of the motor. This method is referred as "Method 2". The built-in automatic meshing feature in JMAG can produce good mesh without jeopardising the convergence of the iteration process while optimising the simulation time. The combinations of mesh size settings including the auto mesh (Case 1) listed in Table 6.1 were examined in this research.

<table>
<thead>
<tr>
<th>Remarks</th>
</tr>
</thead>
<tbody>
<tr>
<td>Case 1</td>
</tr>
<tr>
<td>Case 2</td>
</tr>
<tr>
<td>Case 3</td>
</tr>
<tr>
<td>Case 4</td>
</tr>
<tr>
<td>Case 5</td>
</tr>
<tr>
<td>Case 6</td>
</tr>
<tr>
<td>Case 7</td>
</tr>
<tr>
<td>Case 8</td>
</tr>
</tbody>
</table>
"Method 1" was implemented in Cases 2-5 to improve the back-EMF, cogging torque and axial force waveforms. The half model mesh view of Case 1 and 2 are given in Figure 6.7. Case 2 uses a similar sized stator mesh to Case 1 but a finer airgap mesh. In Cases 3 to 5 a finer mesh size was set on the stator to examine the changes on the simulated results. On the other hand, "Method 2" which uses a smaller stator mesh size was implemented in Cases 6 to 8 for comparison. Figures 6.8 to 6.10 give the back-EMF, flux, cogging torque and axial force waveforms for selected cases (see Appendix B for the results for all the above cases). The simulation times and results for the eight cases are presented in Table 6.2.

![Figure 6.7](image)

**Figure 6.7.** 12S10P machine, JMAG half model mesh view of: a) Case 1 and b) Case 2.

![Figure 6.8](image)

**Figure 6.8.** 12S10P machine, JMAG Case 1 and 2, a) phase back-EMF, b) tooth and yoke flux plots at 1,000rpm.
As it can be seen in Figures 6.8 and 6.9, the waveforms generated in Case 2 were improved significantly compared to Case 1. The generated waveforms of Cases 2 to 5 (see Appendix B) were found generally similar. Nevertheless, the cogging torque and axial force waveforms are improved with the smaller stator mesh sizes. Overall, Case 5 with the smallest mesh size showed the highest accuracy but also required the longest simulation time.

Note that, in Table 6.2, Cases 6 to 8 use a fine mesh size only for the stator. The phase back-EMF waveforms were improved for Cases 7 and 8 as shown in Figure 6.10 as compared to Case 1. Nevertheless, no improvement was observed for the cogging torque and axial force waveforms in Cases 6 and 7 as compared to Case 1 (see Appendix B).
On the other hand, the waveforms of Case 8 were comparable to Case 2. As the simulation time of Case 8 only offers a 10% reduction compared to Case 5, “Method 2” was concluded as not useful for the investigated model.

Table 6.2. FE simulated no-load results for various cases in Table 6.1.

<table>
<thead>
<tr>
<th>Case</th>
<th>Run Time (hours)</th>
<th>Back-EMF (V)</th>
<th>Tooth Flux (mWb)</th>
<th>Yoke Flux (mWb)</th>
<th>Cogging (Nm) max min</th>
<th>Axial Force (N) max min</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>3.6</td>
<td>21.3</td>
<td>0.350</td>
<td>0.222</td>
<td>0.883 -1.301</td>
<td>1141 1225</td>
</tr>
<tr>
<td>2</td>
<td>15.6</td>
<td>20.9</td>
<td>0.352</td>
<td>0.227</td>
<td>0.202 -0.217</td>
<td>1383 1391</td>
</tr>
<tr>
<td>3</td>
<td>20.4</td>
<td>21.1</td>
<td>0.350</td>
<td>0.226</td>
<td>0.212 -0.214</td>
<td>1386 1393</td>
</tr>
<tr>
<td>4</td>
<td>24.7</td>
<td>20.9</td>
<td>0.354</td>
<td>0.229</td>
<td>0.219 -0.261</td>
<td>1409 1399</td>
</tr>
<tr>
<td>5</td>
<td>63.6</td>
<td>20.9</td>
<td>0.354</td>
<td>0.229</td>
<td>0.224 -0.266</td>
<td>1409 1400</td>
</tr>
<tr>
<td>6</td>
<td>6.9</td>
<td>20.9</td>
<td>0.354</td>
<td>0.228</td>
<td>0.760 -1.128</td>
<td>1295 1246</td>
</tr>
<tr>
<td>7</td>
<td>12.9</td>
<td>21.0</td>
<td>0.353</td>
<td>0.227</td>
<td>0.529 -0.433</td>
<td>1361 1339</td>
</tr>
<tr>
<td>8</td>
<td>58.1</td>
<td>20.9</td>
<td>0.354</td>
<td>0.229</td>
<td>0.240 -0.251</td>
<td>1412 1404</td>
</tr>
</tbody>
</table>

In addition, Table 6.3 gives the FE simulated average torque generated with a six-step voltage-source inverter at 546rpm for Cases 1, 2, 4 and 6. From the no-load results, Case 4 was found to closely match the accuracy of Case 5 while requiring about 2.5 times shorter run time (see Table 6.2). Hence it was used as the reference case instead of Case 5 for the comparison under load. Case 6 was included for comparison of the effect of using the “Method 2” meshing on the torque estimation.

Table 6.3. FE simulated loaded (546rpm) results for comparison of average torque for selected cases, under 6-step inverter operation ($V_{DC} = 24V_{dc}$).

<table>
<thead>
<tr>
<th>Case</th>
<th>Average Torque (Nm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1.34</td>
</tr>
<tr>
<td>2</td>
<td>1.29</td>
</tr>
<tr>
<td>4</td>
<td>1.27</td>
</tr>
<tr>
<td>6</td>
<td>1.35</td>
</tr>
</tbody>
</table>
As can be seen in the table above, there is about 4% accuracy improvement with “Method 1” (see Cases 1 and 2) and another 1.5% of accuracy is obtained when the stator mesh size was also reduced (see Cases 2 and 4). Nevertheless, the simulation time of Cases 2 and 4 were significantly higher than Case 1 (see Table 6.2). Case 2 can offers about 40% reduction in simulation time while sacrificing 1.5% of the accuracy compared to Case 4. The increase of 4% accuracy compared to the simulation times in Cases 1 and 2 was not justifiable considering the high number of different conditions to be simulated, which will be discussed later in Chapters 7 and 8.

Reducing the mesh size only (“Method 2”) was found not to be as useful as in Case 2 considering the simulation time. This was deduced from the fact that Case 6 with mesh size of \(3 \text{mm}\) is the only case with simulation time less than Case 2 (about 50%) but showed no improvement (see Cases 4 and 6) in the accuracy of the generated average torque.

As a result, in order to optimise the computation time, Case 2 is chosen to be utilised mainly for generating cogging torque and axial force waveforms and Case 1 is chosen as the main simulation model for the designs in Chapters 7 and 8.

### 6.2.2 Simulation Time and File Size

In this section, the simulation times and file sizes obtained when selecting mesh sizes which generate acceptable accuracy for different models were compared. The accuracy level of the model was tested based on producing smooth flux distributions and induced voltage waveforms.

Table 6.4 summarises the simulation time, number of elements and solution file size for the different models previously shown in Figure 6.1. Note that, ”auto-mesh” is referring to the built-in automatic mesh sizing selection tool and the ”manual mesh” is referring to manually setting the mesh sizing (see Case 2 in Table 6.2). Also, the models were set to calculate the instantaneous torque and tooth flux during the simulation.

Even though the 110mm 12S10P half model with the auto-mesh setting showed the lowest solution time and file size, the generated waveforms were not acceptable as shown in Section 6.2.1. Based on the optimisation method shown, the half model required the longest execution time and produced the largest file size compared to the other models. The full model of the 32mm 3S2P machine resulted with more elements compared to the 110mm 12S4P machine quarter model using the auto-mesh feature.
Therefore, the quarter model required the least execution time and generated the smallest file size.

<table>
<thead>
<tr>
<th>Model</th>
<th>Simulation Time (hours)</th>
<th>Number of Elements</th>
<th>File Size (GB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>No-load</td>
<td>Load</td>
<td></td>
</tr>
<tr>
<td>32mm 3S2P full model</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>(auto-mesh)</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>- W/O back-iron</td>
<td>3.8</td>
<td>5.1</td>
<td>82,600</td>
</tr>
<tr>
<td>- W/ back-iron</td>
<td>7.6</td>
<td>8.2</td>
<td>86,300</td>
</tr>
<tr>
<td>110mm 12S4P quarter model</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>(auto-mesh)</td>
<td>3.8</td>
<td>4.8</td>
<td>60,900</td>
</tr>
<tr>
<td>110mm 12S10P half model</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>(auto-mesh)</td>
<td>3.8</td>
<td>4.6</td>
<td>42,900</td>
</tr>
<tr>
<td>(manual mesh, Case 2)</td>
<td>15.6</td>
<td>15.8</td>
<td>184,100</td>
</tr>
</tbody>
</table>
6.3 Analysis Procedures

The analysis procedures described in the following sections were based on the 12 slot stator with 12mm slot depth designed for four poles (12S4P). Due to the available resources, the 12mm slot depth stator was also used for the ten-pole machine (12S10P).

The parameter and performance measurement procedures described in Section 7.3 were conducted. High bandwidth current and voltage transducers (Hameg HZ56 current probe and Yokogawa Model 700925 differential voltage probe), a digital oscilloscope (Tektronix TDS 340A) and a three-phase power analyser (Voltech PM3000ACE) were utilised to obtain the test data. The measured waveforms were presented and compared to the FE simulated results in the following sections.

6.3.1 Machine Parameters

The open-circuit test was simulated by rotating the rotor with no inverter connected to the three-phase winding and hence with zero excitation. The JMAG coupled-circuit diagram is shown in Figure 6.11. In this section, the flux, back-EMF, cogging torque, axial force and iron loss were studied.

![Figure 6.11. Open-circuit condition circuit diagram in JMAG simulation.](image)
Flux and Back-EMF

In JMAG, the tooth flux is available directly as a result display option (FEM coil flux). However, the yoke flux can be calculated by selecting the yoke cross-sectional face on the stator and defining it as a variable. In addition, the back-EMF waveform was obtained by setting the three-phase winding (FEM coil) condition in the coupled-circuit.

Figures 6.12 and 6.13 show the simulated and measured magnetic flux density waveforms in the stator teeth and yoke under open-circuit conditions for the 12S4P and 12S10P machines respectively. The calculated magnetic flux density distribution can also be presented in vector and contour views as shown in Figure 6.14 for the 12S4P case. In addition, Figure 6.15 shows the measured and simulated back-EMF waveforms of two machine configurations under open-circuit condition at 1,000 rpm.

As can be seen from the figures, there is a good correspondence between the measured and simulated waveforms. This gives confidence in the accuracy of the 3D FEA approach and automatic mesh generation method. In addition, it was observed that 12S10P machine produces sinusoidal flux and back-EMF waveforms as expected due to its fractional-slot structure which reduces the high back-EMF harmonics.

![Tooth flux density](a) Tooth flux density

![Yoke flux density](b) Yoke flux density

Figure 6.12. 12S4P machine, FE simulated and measured open-circuit tooth and yoke flux density waveforms at 1,000 rpm.
Chapter 6

3D Finite Element Modelling and Analysis

Figure 6.13. 12S10P machine, FE simulated and measured open-circuit tooth and yoke flux density waveforms at 1,000 rpm.

Figure 6.14. 12S4P machine, a) magnetic flux vector and b) contour plots.

Figure 6.15. FE simulated and measured phase back-EMF waveforms at 1,000 rpm.
Cogging Torque and Axial Force

The cogging torque and axial force were obtained in the simulation by setting the torque and electromagnetic force calculation condition. Under the open-circuit condition, the torque generated is the cogging torque (see Figure 6.16). In addition, the electromagnetic force condition includes calculation of forces in the x, y and z directions. The z direction was set to be the axial direction of the AFPM model and hence the z-axis force represents the axial force as given in Figure 6.17 for the two machine configurations.

![Cogging Torque and Axial Force Graphs](image)

**Figure 6.16.** FE simulated cogging torque as a function of electrical angle at 1,000rpm. The upper and lower range of peak values of measured cogging torque is shown by two horizontal lines.

![Axial Force Waveforms Graphs](image)

**Figure 6.17.** FE simulated axial force waveforms at 1,000rpm.
The peak cogging torque was measured using the method described later in Section 7.3.1. A number of tests were conducted at different initial rotor positions and the upper and lower measured values are shown by horizontal lines in Figure 6.16. It was observed that the measured upper peak values were comparable to the predicted values for 12S4P case. For the 12S10P design the calculated cogging torque is about ten times smaller than that calculated for the 12S4P design. This was expected as fractional-slot designs generally have low cogging torque. The measured upper peak values of 12S10P machine are about double the calculated values. This could be due to assembly and magnetisation tolerances. It can be noted that, the cogging torque can be reduced by utilising a double-sided stator design (a stator on each side of the disc rotor) configuration where the stators can be offset (skewed) with respect to one another.

Figure 6.17 shows that the single-sided configuration has a large average axial force of about 1600\(N\) for the 12S4P design and about 1400\(N\) for the 12S10P design. It can be emphasised here that such large axial force has the potential to produce large bearing losses.

**Iron Loss**

In JMAG, the iron loss was calculated using the measured iron loss characteristics of the AMM material in Section 3.2.2. It should be noted that unless stated otherwise, the measured iron loss curves were utilised in this thesis. This is especially important in AMM based machine where the measured iron loss includes additional loss from the interlaminar eddy-current loss as discussed in Section 3.5.

A time-stepping FEA solution was performed, and the flux density in each element in the model as a function of time was calculated. This was used to calculate the iron loss in each element, which was then integrated over the entire model volume to find the total iron loss. The calculated iron loss is the sum of both the hysteresis loss and eddy-current loss components as shown in Figures 6.18(a) and 6.19(a). As can be seen in the figures, the 12S4P and 12S10P machines showed similar values of iron losses at same low operating speed. This is due to the fact that 12S10P design has higher frequency and the 12S4P design has higher peak tooth and yoke flux densities. In addition, the flux waveforms of the 12S4P machine have higher harmonics compared to the 12S10P machine as shown previously in Figures 6.12 and 6.13. However, the 12S10P machine displayed higher iron loss as speed increases. Further comparisons about the losses will be presented later in Section 7.4.2.
Figures 6.18(b) and 6.19(b) show the measured open-circuit loss and simulated iron loss versus speed characteristic. It is observed that the simulated values were about 20 times smaller than the measured values. This could be due to the test machine’s windage, bearing and rotor eddy-current losses. This is also investigated further in Chapter 8.

In addition, the FE simulation tool also can generate contour plots for the hysteresis, eddy-current and total iron loss density ($W/m^3$). This provides clear insights about the loss distribution in the stator. From the total iron loss density contour plot shown...
in Figure 6.20, it can be clearly seen that the loss is concentrated on the stator teeth surface. At different rotor position, the loss in some teeth is higher than the others (see Figure 6.20(b)). In addition, as discussed previously the 12S4P machine showed higher iron loss in the yoke area due to higher peak flux density.

![Figure 6.20. FE simulated iron loss density contour plots at 1,000 rpm.](image)

(a) 12S4P machine, (W/m³)  
(b) 12S10P machine, (W/m³)

### 6.3.2 Time-Stepping Simulation of Inverter Operation

In order to calculate the machine performance with an inverter, a coupled-circuit time-stepping analysis was used. The circuit model uses a six-step voltage-source inverter and star-connected phase windings as shown in Figure 6.21. The pulse-amplitude modulation (PAM) inverter component was used to control the switching commutations which was set to align with the back-EMF waveform. Different loading conditions were simulated by changing the voltage level and rotation speed.

![Figure 6.21. Loaded condition circuit model in JMAG simulation.](image)
Inductance

The inductance of the motor was calculated by connecting a three-phase current source to the motor windings. The inductance calculation tool calculates the inductance in both the $d$-axis (magnet axis) and the $q$-axis from the standard $d$- and $q$-axis voltage equations as given in Equation 6.1 [171]. The tool finds the three $q$-axis voltages ($V_{q0}, V_{q1}, V_{q2}$) corresponding to three $d$-axis currents ($I_{d0}, I_{d1}, I_{d2}$). It then finds the $d$-axis inductance ($L_d$) using Equations 6.2 to 6.4. The $q$-axis inductance ($L_q$) is found using the $d$-axis voltage $V_d$ corresponding to a $q$-axis current $I_q$ (see Equation 6.5). The numerical values of current amplitude, current phase, $d$- and $q$-axis inductances are generated. Due to the axial surface magnet configuration, the calculated inductances in the two axes are expected to be similar.

\[
\begin{pmatrix} V_d \\ V_q \end{pmatrix} = \begin{pmatrix} R & -\omega L_q \\ \omega L_d & R \end{pmatrix} \begin{pmatrix} I_d \\ I_q \end{pmatrix} + \begin{pmatrix} 0 \\ \omega \psi_a \end{pmatrix}
\]

\[
L_{d1} = \frac{V_{q0} - V_{q1}}{\omega (I_{d0} - I_{d1})}
\]

\[
L_{d2} = \frac{V_{q0} - V_{q2}}{\omega (I_{d0} - I_{d2})}
\]

\[
L_d = \frac{L_{d1} + L_{d2}}{2}
\]

\[
L_q = \frac{-V_d}{\omega I_q}
\]

where $R$: phase resistance and $\psi_a$: magnet flux-linkage.

Figures 6.22 and 6.23 show the comparison between the simulated and measured inductances, with and without the rotor in place for the 12S4P and 12S10P machines respectively. Similar $L_d$ and $L_q$ values were simulated which was expected for the surface-mounted configuration. It was found that the presence of the rotor increases the inductance by about 25% for 12S4P and 40% for 12S10P configuration.

The simulated results show a good correspondence with the measured values for the 12S4P machines, but show a 15% discrepancy in the case of 12S10P machine. This could be due to extra leakage flux and possible variations of inductance with rotor position.
Voltage and Current

The machine voltage and current waveforms were obtained by setting the three-phase winding (FEM coil) condition in the coupled-circuit as shown in Figure 6.21. In addition, the current commutation phase angle was set to obtain the maximum output torque where the voltage and current waveforms are in phase.
Figures 6.24 and 6.25 show the simulated and measured phase voltage and phase current waveforms for the 12S4P and 12S10P machines respectively. The simulation results were in reasonable agreement with the measured waveforms. The small discrepancy in the current waveforms is likely to be due to slight differences in the practical commutation timing or errors in the measured inductance.

**Figure 6.24.** 12S4P machine, FE simulated and measured phase voltage and current waveforms at 1018rpm and DC link voltage of 24V_{dc}.

**Figure 6.25.** 12S10P machine, FE simulated and measured phase voltage and current waveforms at 546rpm and DC link voltage of 24V_{dc}.
Torque

The instantaneous output torque under load was calculated (see Figures 6.26(a) and 6.27(a)) by setting the torque calculation condition (nodal force method) in JMAG. The torque contribution of each node is equal to the multiplication of the electromagnetic force (circumferential direction component) that acts on the rotor by the distance from the rotating axis to the nodal point. Then, the total torque is obtained from the sum of the nodal torque components over the rotor region. The average torque can be obtained by averaging the steady-state instantaneous torque waveform (see Figures 6.26(b) and 6.27(b)).

![Figure 6.26. 12S4P machine, a) FE simulated instantaneous electromagnetic torque, b) measured and FE simulated average torque at 1018 rpm and DC link voltage of 24 Vdc.](image)

The experimental results were also shown in Figures 6.26(b) and 6.27(b). The simulation results are in reasonable agreement with the measured curves for the 12S4P machine but have very good correspondence with the 12S10P machine. The discrepancies are likely to be due to the discrepancy of practical commutation timing.

Eddy-Current Loss

The eddy-current loss is calculated by setting a finite electric conductivity of the rotor materials (i.e. magnet and back-iron). If the skin depth is much smaller than the magnet size, the skin depth is included in the mesh settings option in JMAG.
Figures 6.27 and 6.29 show the generated “Joule Loss” which gives the eddy-current losses of one magnet pole and its associated rotor back-iron for the 12S4P and 12S10P machines. The eddy-current waveforms simulated were less regular for the 12S10P machine. Section 8.3.1 presents a more detailed investigation on eddy-current loss in the rotor.

Figure 6.27. 12S10P machine, a) FE simulated instantaneous electromagnetic torque, b) measured and FE simulated average torque at 546rpm and DC link voltage of 24Vdc.

Figure 6.28. 12S4P machine, FE simulated magnet and rotor back-iron eddy-current loss at 546rpm, loaded.
Figure 6.29. 12S10P machine, FE simulated magnet and rotor back-iron eddy-current loss at 546rpm, loaded.

The contour plots of magnet and rotor back-iron eddy-current loss density for the 12S4P and 12S10P machines are given in Figures 6.30 and 6.31. From the figures, it can be seen that the rotor back-iron loss is concentrated on the surface where the magnets are glued on. The loss in the magnets varies with rotor position.

Figure 6.30. 12S4P machine, FE simulated magnet and rotor back-iron eddy-current loss density contour plots at 546rpm, loaded.
Efficiency

Figure 6.32 shows the measured and simulated efficiency (without bearing, windage and eddy-current losses) as a function of load torque of the two motors under test. It should be noted that the measured efficiency was based on the output torque estimated using a dc motor as will be described in Section 7.3.2 due to the absence of a torque transducer. The maximum measured efficiency was 65% at a speed of 1018 rpm for the 12S4P machine and 72% for the 12S10P machine at a speed of 640 rpm. The FE calculated efficiency was significantly higher in both motor configurations, as the bearing, windage and eddy-current losses (see Section 8.1) were not taken into consideration at this stage.

For demonstration purposes, the bearing, windage and eddy-current losses were included to calculate the simulated efficiency and is plotted in Figure 6.33. The FE calculated efficiency closely matched the measured values.
Figure 6.32. Measured and simulated efficiency (without bearing, windage and eddy-current losses) versus output torque characteristics at DC link voltage of $24V_{dc}$.

Figure 6.33. Measured and simulated efficiency (with bearing, windage and eddy-current losses) versus output torque characteristics at DC link voltage of $24V_{dc}$. 
6.4 Operation at High Current

At high current operation of the AFPM machine, there is a risk in demagnetising the magnets and also saturating the magnetic materials (AMM) of the stator core. In this section, the effect is investigated with 3D FEA.

6.4.1 Demagnetisation

Figure 6.34 shows the FE airgap axial flux density plot as a function of circumferential position at a temperature of 140°C (magnet limitation) for the open-circuit case and for currents of 4.5\(A_{pk}\), 9\(A_{pk}\), 18\(A_{pk}\) (demagnetisation level) and 36\(A_{pk}\) (high demagnetisation) based on the B-H curve discussed previously in Section 5.12. Note that, the airgap magnet flux waveform was lowered down by the negative \(d\)-axis current.

![Figure 6.34](image)

**Figure 6.34.** FE calculated airgap axial flux density as a function of circumferential position at various negative \(d\)-axis current values at 140°C.

6.4.2 AMM Saturation Effect

In order to investigate the effect of saturation on the output torque, the output torque simulation studies were carried out at different current densities. This was conducted using a given value of dc current flowing in one machine terminal and exiting the other two terminals as shown in Figure 6.35. Then, the maximum electromagnetic torque corresponding to the rotor angle was obtained which approximates the average output torque as the cogging torque is low in the 12S10P design.
First, the output torque versus slot current density curves with both a linear material of 600,000 relative permeability (AMM material) and with the measured AMM B-H characteristics were calculated and compared (see Figure 6.36(a)). As expected, there is no saturation with the linear material and therefore the torque increases linearly with current density. In addition, there is a good match at low current densities between the two materials which gives confidence in the accuracy of the measured AMM B-H curve imported into JMAG. The torque curve starts to bend over for current densities greater than $100 \text{A/mm}^2$ showing the effect of saturation in the AMM. The corresponding peak tooth flux density is $1.35T$. This is consistent with the effective saturation flux density (stacking factor of $0.89 \times 1.5T$) of the AMM laminations. Nevertheless, $100 \text{A/mm}^2$ is significantly higher than the thermal limits of the winding and hence would not be practical. Therefore within the thermal limits of the winding, the motor design would not be expected to be affected by the saturation of the AMM material.

Figure 6.36. 12S10P machine, FE simulated output torque for linear magnetic material and AMM versus current density.

(a) AMM and linear material (SD=12mm)  
(b) AMM, slot depth variation
Ignoring the winding thermal limits the output torque versus current density for different slot depths were simulated and shown in Figure 6.36(b). As discussed previously, there is a linear relation between the peak output torque and the slot area. A larger slot area would produce a higher peak torque capability but saturation occurs at a lower current density value due to larger slot area.

In addition, Figure 6.37 gives the simulated peak torque for different slot depths and slot widths within the linear region in terms of the number of slot ampere-turns. As shown previously, changing the slot depth is more effective than changing the slot width in obtaining a larger torque. Nevertheless, the magnets are also subject to a demagnetisation risk at high currents especially at high temperatures. The demagnetisation value obtained from Section 5.12 was also plotted in Figure 6.37. This acts as another limit on the peak output torque.

![Figure 6.37](image)

(a) Changing slot depth, (SW=8mm)  
(b) Changing slot width, (SD=12mm)

**Figure 6.37.** 12S10P machine, FE calculated peak torque versus ampere-turns at various values of slot depth and slot width.
6.5 Conclusions

Three dimensional FEA is a numerical method that is able to accurately analyse the complex 3D electromagnetic fields in machines including the leakage and end-winding fields. Hence, it offers high accuracy parameter and performance predictions of the machine being modelled and is required for axial-field PM machines due to their inherent 3D geometry. Nevertheless, the high level of accuracy in simulation requires long computation time which is not favorable in the initial machine design process. It is more often utilised to verify the analytical model.

The primary focus of this chapter is to describe the utilisation of the JMAG 3D FEA package to model AFPM machines and hence to predict the parameter and performance characteristics of these machines. In this chapter, the mesh size selection and optimisation process based on computation time were proposed and described in detail. A comparison of the simulation times, file sizes and resultant accuracy of different models was performed. It was found that mesh optimisation is not required for the quarter model of 12S4P machine while some optimisation is performed for the half model of 12S10P machine.

Detailed demonstrations to obtain the common parameters of induced back-EMF, cogging torque, iron loss, output torque and inductance were presented. In addition, the tooth and yoke flux distributions and the time stepping voltage and current waveforms which are not commonly reported in publications were also investigated. In addition, 3D FEA was utilised to predict and analyse iron and eddy-current losses, demagnetisation and saturation effects. Overall, good agreement between simulation and experiment results was obtained. The simulation results showed that the saturation of the AMM material has no significant effect on the torque production of the 12S10P machine. In addition, it was found that it is more effective to change the slot depth than slot width of the 12S10P machine to obtain a larger torque.

This chapter also demonstrates the validity of the 3D FEA approach against comprehensive experimental results. The motor parameters and its performance were measured and compared with the 3D FEA predictions. The validation of the use of 3D FEA to model AFPM machines is the first step in optimising the machine geometry to improve its performance.
Chapter 7

110mm Machine 3D FE Analysis and Experimental Results

This chapter considers a larger size (110mm) AFPM motor design utilising cut AMM material. The stator was constructed and the characteristics of the motor configuration was investigated experimentally and by simulation utilising 3D FEA. The details of the comprehensive experimental and analysis results of the prototype AMM machine are presented in this chapter. This includes examination of the machine’s parameters, airgap length, pole configuration and magnet material. In addition, identical stators of different materials were constructed for a direct performance comparison.
7.1 3D FE Analysis of Machine Design Variables

In Section 6.3, it was shown that there is a good agreement between 3D FEA simulation and experimental results obtained for the 12S4P and 12S10P machines. Hence, there is a high degree of confidence in using 3D FEA for simulating detailed characteristics of AFPM machines. In this section, the models created were utilised to analyse and compare the effect of design variations (i.e. number of poles, magnet and stator materials) of a larger size machine (110mm diameter) using the 12S10P AMM stator with 12mm slot depth (SD) and sintered magnets as the baseline design.

7.1.1 12mm versus 17mm Slot Depth

As discussed in Chapter 5 for the selected core size and AMM ribbon, the 17mm slot depth AMM stator core design was chosen as an optimum depth. Nevertheless, the practical machine was not available for testing. Therefore, the 12mm slot depth stator designed for the 12S4P machine was utilised in the verification study. However, the 17mm slot depth SMC and SI stators were constructed and were available for testing. In addition, 3D FEA models of the machines were developed to examine the effect of the 5mm variation in slot depth.

It was found that the waveform shape and magnitude of the flux densities, back-EMF, cogging torque and axial force waveforms obtained at open-circuit conditions with the 12mm machine were comparable to the 17mm slot depth machine (see Figures 7.6, 7.9(a) and 7.11(b)). As shown in Figure 7.1, the main difference between the 12mm and 17mm designs is the smaller yoke flux density due to the large yoke area (smaller slot depth).

The simulation waveforms generated under the loaded conditions are given in Figures 7.2 to 7.4. As obtained at open-circuit conditions, similar tooth flux densities and smaller peak yoke flux density (due to the larger yoke area) were calculated with the 12mm slot depth stator (see Figure 7.2(b)). Similar phase voltage and current waveforms were generated as expected with the identical teeth shape, number of turns and the same coupled-circuit settings as shown in Figure 7.3. Also, both slot depth stators have similar axial force and cogging torque due to their identical tooth shape and magnet remanence. Although small differences in the waveform shape is observed in Figure 7.4(a), the average torque is similar which is due to similar current levels in both cases.
Nevertheless, there is a significant difference in the stator iron loss as shown in Figure 7.4(b). As it can be seen in the figure, there is an approximate 10% drop in iron loss in the 12mm slot depth stator even though it has a greater iron volume. This is due to the smaller peak yoke flux density with the larger yoke area and the fact that the iron loss is proportional to the flux density squared. On the other hand, the increase in slot depth has only a small effect on the magnet and rotor eddy-current loss as shown in Figure 7.5. Note that, the resistivity used for the sintered magnet is $1.5 \times 10^{-6} \Omega m$ and for the rotor back-iron is $2.22 \times 10^{-7} \Omega m$. 
Figure 7.3. 12S10P machine loaded, FE simulated phase back-EMF and current waveforms with 12\(\text{mm}\) and 17\(\text{mm}\) slot depths at 546\(\text{rpm}\).

Figure 7.4. 12S10P machine, FE simulated output torque waveform at 546\(\text{rpm}\) and open-circuit iron loss with 12\(\text{mm}\) and 17\(\text{mm}\) slot depths.
Table 7.1 gives the measured resistance and inductance for the 12\(\text{mm}\) and 17\(\text{mm}\) slot depth AMM and SI stators which were built and tested. As expected, a lower resistance (35\% drop) was obtained with the 17\(\text{mm}\) slot depth stator due to the thicker wire. However, this design increased the inductance by 5\% due to the deeper slot depth resulting in greater slot leakage.

Table 7.1. 12S10P machine, measured resistance and inductance for 12\(\text{mm}\) and 17\(\text{mm}\) slot depths

<table>
<thead>
<tr>
<th>Slot Depth ((\text{mm}))</th>
<th>Resistance ((\Omega))</th>
<th>Inductance (without rotor) ((\text{mH}))</th>
</tr>
</thead>
<tbody>
<tr>
<td>12 (AMM)</td>
<td>0.46</td>
<td>0.49</td>
</tr>
<tr>
<td>17 (SI)</td>
<td>0.30</td>
<td>0.56</td>
</tr>
</tbody>
</table>

The smaller resistance results in lower copper loss and since this is the dominant loss at low speed, it is a more influential factor affecting the efficiency compared to iron loss under low speed operating conditions. From Table 7.2, a 6\% higher efficiency was predicted for the 17\(\text{mm}\) stator compared to the 12\(\text{mm}\) stator. Therefore, it could be concluded that the 12\(\text{mm}\) stator would still provide a reasonable experimental comparison with the 17\(\text{mm}\) SI and SMC based stators which will be investigated later in Section 7.4.4.
Table 7.2. 12S10P machine, FE simulated performance for 12mm (AMM) and 17mm (SI) slot depths at 546 rpm.

<table>
<thead>
<tr>
<th>Slot Depth (mm)</th>
<th>Output Torque (Nm)</th>
<th>Stator Iron Loss (W)</th>
<th>Copper Loss (W)</th>
<th>Rotor Eddy-current Loss (W)</th>
<th>Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>12</td>
<td>1.35</td>
<td>0.58</td>
<td>17.5</td>
<td>3.40</td>
<td>78</td>
</tr>
<tr>
<td>17</td>
<td>1.37</td>
<td>0.67</td>
<td>11.1</td>
<td>3.14</td>
<td>84</td>
</tr>
</tbody>
</table>

### 7.1.2 Tooth and Yoke Flux Densities

Figure 7.6 gives the tooth and yoke flux density waveforms simulated with the AMM 12S10P machine with 17mm slot depth under open-circuit conditions. The results from SI and SMC machines with 17mm slot depth are also included in the figure. During these studies, the material types were changed but the mesh size was kept constant. As can be seen from the figures, the waveforms were comparable for all three materials.

![FE tooth flux density](a)

![FE yoke flux density](b)

**Figure 7.6.** 12S10P machine, FE simulated open-circuit tooth and yoke flux density waveforms of AMM, SI and SMC at 1,000 rpm, with 17mm slot depth

Figure 7.7 shows that the tooth and yoke flux waveforms generated with the 12S10P machine (12mm slot depth) are more sinusoidal compared to the 12S4P machine (12mm slot depth). The flux waveforms generated with the 4P rotor have a wider flat peak in the tooth flux density waveform (see Figure 7.7(a)) and a sharper peak in the yoke waveforms.
flux density waveform (see Figure 7.7(b)). Furthermore, for the 4P rotor, the peak flux density values were higher by about 1.3 times for the tooth flux density and about 3 times for the yoke flux density.

![Figure 7.7](image-url)  
(a) FE tooth flux density  
(b) FE yoke flux density

Based on the 12mm slot depth AMM stator and the same number of turns as the sintered magnet design, a 6mm thickness bonded magnet design which is twice the thickness of the sintered magnet was chosen due to ease of implementation. The low remanent flux density of the bonded magnets means that it was unable to match the airgap flux density of the sintered magnet design at the same airgap length. Figure 7.8 shows the flux-linkage waveforms for both magnet types at an airgap length of 1mm. As expected, the flux-linkage of the bonded magnet design with the same number of turns is lower than the sintered magnets design. As shown in Figure 7.8(a), the measured tooth flux density using bonded magnets was also included in the plots which matches closely.

In order to compare the performance characteristics, a similar phase voltage and hence flux-linkage values were adopted. This can be implemented by increasing the number of stator turns from 24 to 34 for the bonded magnet design. Nevertheless, only the 24 turn AMM stator was available for testing. Hence, the airgap length for the sintered magnet design was increased to 3mm to match the back-EMF constant of the bonded magnet design at a 1mm airgap length. As shown, the FE simulation results with the 3mm airgap was also included in Figure 7.8. It can be seen that the resultant tooth
and yoke flux densities are slightly lower but comparable to the values of the bonded magnet design.

![Graphs showing FE tooth flux density and yoke flux density](image)

**Figure 7.8.** 12S10P machine, FE simulated open-circuit tooth and yoke flux density waveforms of sintered and bonded magnets at 1,000 rpm. An airgap of 1 mm was used unless otherwise stated.

### 7.1.3 Phase Back-EMF

Figure 7.9(a) gives the phase back-EMF generated at open-circuit conditions for the AMM, SI and SMC based machines. As expected, the phase back-EMF waveforms were comparable for the stators with the same number of turns, airgap length and identical tooth designs. The tooth flux waveforms generated with the 4P rotor have a wide flat-top (see Figure 7.7(a)). As a result, the corresponding back-EMF profile has a wider zero level and narrower flat-top as shown in Figure 7.9(b). On the other hand, a much more sinusoidal back-EMF waveform is generated with the 10P rotor.

Figure 7.10 shows the FE simulated phase back-EMF and flux-linkage waveforms for both magnet types at an airgap length of 1 mm and a speed of 1,000 rpm. As it can be seen the waveforms match well when using an increased number of turns for the bonded magnet design. In addition, the measured bonded magnet result also matches closely as shown in 7.10(a). Note that changing the number of turns only changes the flux-linkage and hence back-EMF waveform, but not the flux density. As for the sintered magnet design with a 3 mm airgap, the waveforms are slightly lower but comparable to the bonded magnet design.
Figure 7.9. FE simulated phase back-EMF waveforms: a) 12S10P AMM, SI and SMC machines and b) AMM 12S10P and 12S4P machines at 1,000 rpm.

Figure 7.10. 12S10P machine, FE simulated phase back-EMF and flux-linkage waveforms of AMM stator with sintered and bonded magnets at 1,000 rpm.

Table 7.3 summarises the analytically calculated, FE simulated and measured phase back-EMF constants. The analytical values were calculated using the induced voltage equation (see Equation 5.68). As listed, the analytical values are reasonably close to the measured values (with a largest discrepancy of 35% for 10P bonded magnet design). On the other hand, the FE simulated values display a good agreement with all the values, within 6% of the measured results.
Table 7.3. Analytically calculated, FE simulated and measured back-EMF constant results.

<table>
<thead>
<tr>
<th>Material</th>
<th>Analytical (V/rad/s)</th>
<th>FE Simulated (V/rad/s)</th>
<th>Measured (V/rad/s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>AMM</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4P, sintered</td>
<td>0.104</td>
<td>0.0968</td>
<td>0.992</td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.177</td>
<td>0.145</td>
<td>0.144</td>
</tr>
<tr>
<td>- 3mm airgap</td>
<td>0.121</td>
<td>0.0889</td>
<td>0.0943</td>
</tr>
<tr>
<td>10P, bonded</td>
<td>0.126</td>
<td>0.0981</td>
<td>0.0936</td>
</tr>
<tr>
<td>10P, bonded (34 turns)</td>
<td>0.179</td>
<td>0.143</td>
<td>-</td>
</tr>
<tr>
<td>SI</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.177</td>
<td>0.147</td>
<td>0.145</td>
</tr>
<tr>
<td>SMC</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.177</td>
<td>0.139</td>
<td>0.143</td>
</tr>
</tbody>
</table>

7.1.4 Inductance Values

The analytically calculated, FE simulated and measured inductances are given in Table 7.4 for 12S10P and 12S4P machines with 12mm and 17mm slot depths, with and without sintered and bonded magnet rotors. Due to the surface PM machine topology implemented in the design, the machine shows little saliency between the \( d \)- and \( q \)-axis. The results show that the FE simulated values are within 15% of the measured values. In addition, the presence of the rotor increases the inductance by an average of 30%.

The measured inductance value of the 10P design is about 2 times higher compared to the 4P design without rotor and about 1.3 times with rotor. This reduction could be due to the larger magnetising inductance with the 4 pole rotor. In addition, the AMM and SI designs showed similar inductance values without the rotor. Nevertheless, it was about 6% lower in the SMC design, which is caused by the smaller amplitude of flux-linkage waveform in the SMC core as it has lower permeability (see Figure 7.6). Similar results were observed for the inductance with the rotor.
Table 7.4. Analytically calculated, FE simulated and measured inductance $L$.

<table>
<thead>
<tr>
<th>Material</th>
<th>Anal $L_q$ (mH)</th>
<th>Anal $L_d$ (mH)</th>
<th>FE $L_q$ (mH)</th>
<th>FE $L_d$ (mH)</th>
<th>Meas $L$ (mH)</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Without rotor</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>AMM (SD12mm)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4P, sintered</td>
<td>-</td>
<td>-</td>
<td>0.24</td>
<td>0.24</td>
<td>0.22</td>
</tr>
<tr>
<td>10P, sintered</td>
<td>-</td>
<td>-</td>
<td>0.45</td>
<td>0.45</td>
<td>0.49</td>
</tr>
<tr>
<td>10P, bonded (34 turns)</td>
<td>-</td>
<td>-</td>
<td>0.84</td>
<td>0.84</td>
<td>-</td>
</tr>
<tr>
<td>SI (SD17mm)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10P, sintered</td>
<td>-</td>
<td>-</td>
<td>0.51</td>
<td>0.51</td>
<td>0.56</td>
</tr>
<tr>
<td>SMC (SD17mm)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10P, sintered</td>
<td>-</td>
<td>-</td>
<td>0.48</td>
<td>0.48</td>
<td>0.52</td>
</tr>
<tr>
<td><strong>With rotor</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>AMM (SD12mm)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4P, sintered</td>
<td>0.68</td>
<td>0.67</td>
<td>0.53</td>
<td>0.53</td>
<td>0.55</td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.49</td>
<td>0.48</td>
<td>0.62</td>
<td>0.62</td>
<td>0.71</td>
</tr>
<tr>
<td>- 3mm airgap</td>
<td>0.47</td>
<td>0.46</td>
<td>0.50</td>
<td>0.50</td>
<td>0.59</td>
</tr>
<tr>
<td>10P, bonded</td>
<td>0.41</td>
<td>0.38</td>
<td>0.53</td>
<td>0.53</td>
<td>0.62</td>
</tr>
<tr>
<td>10P, bonded (34 turns)</td>
<td>0.82</td>
<td>0.76</td>
<td>1.05</td>
<td>1.04</td>
<td>-</td>
</tr>
<tr>
<td>SI (SD17mm)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.55</td>
<td>0.51</td>
<td>0.69</td>
<td>0.69</td>
<td>0.72</td>
</tr>
<tr>
<td>SMC (SD17mm)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.54</td>
<td>0.53</td>
<td>0.66</td>
<td>0.66</td>
<td>0.62</td>
</tr>
</tbody>
</table>

The analytically estimated inductances for the 4P design are about 28% higher than the measured values while 28% lower was found in the 10P design. This is likely to be due to large errors in calculating the leakage inductance. In addition, the FE simulated and measured values of 4P design is smaller than the 10P design. However, based on the magnetising inductance equation in Section 5.7, it is expected that the inductance in the 4P design to be about 1.4 times higher than the 10P design, which is due to the smaller pole number and winding factor. This is again likely due to the errors in the leakage inductance estimation.
The lower value of the predicted inductance (15%) with the bonded magnet design (compared to the sintered magnet design) could be due to the increased leakage as the magnet thickness is doubled. On the other hand, the FE simulated inductances are found within an average of \( \pm 5\% \) of the measured values. The bonded magnet design showed the largest difference of about 17% lower compared to the measured values. In addition, there is an increase in inductance for the bonded magnet design with higher number of turns. As known, this is due to the fact that the inductance is proportional to the number of turn squared.

Table 7.5 gives the analytically calculated components of the inductance based on the equations in Section 5.7. Due to the 3D axial structure and the patch mesh method (see Section 6.2.1) used in the FE simulation, the FE leakage inductances were not available. The dominant leakage inductance was the slot leakage component. For the 4P configuration, the slot leakage was about 1.6 times smaller while the end-windings leakage was 2.5 times higher compared to the 10P configuration. Increasing the number of turns from 24 to 34 has the effect of doubling all the inductances. Increasing the airgap length increases the tooth tip leakage but decreases the magnetising inductance. In addition, as expected, the difference in slot depth only changes the slot leakage inductance.

### Table 7.5. Analytically calculated armature reaction \( q \)-axis inductance \( L_{aq} \), slot leakage \( L_{1s} \), end-windings leakage \( L_{1ein}, L_{1eout} \), tooth tip leakage \( L_{1tt} \) inductances.

<table>
<thead>
<tr>
<th>Material</th>
<th>( L_{aq} ) ((mH))</th>
<th>( L_1 )</th>
<th>( L_{total} ) ((mH))</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>( L_{1s} )</td>
<td>( L_{1ein} )</td>
</tr>
<tr>
<td>AMM (SD12\text{mm})</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4P, sintered</td>
<td>0.43</td>
<td>0.085</td>
<td>0.039</td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.26</td>
<td>0.14</td>
<td>0.015</td>
</tr>
<tr>
<td>- 3\text{mm} airgap</td>
<td>0.17</td>
<td>0.14</td>
<td>0.015</td>
</tr>
<tr>
<td>10P, bonded</td>
<td>0.17</td>
<td>0.14</td>
<td>0.015</td>
</tr>
<tr>
<td>10P, bonded (34 turns)</td>
<td>0.35</td>
<td>0.28</td>
<td>0.031</td>
</tr>
<tr>
<td>SI (SD17\text{mm})</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.26</td>
<td>0.20</td>
<td>0.015</td>
</tr>
</tbody>
</table>
Chapter 7  

110mm Machine 3D FE Analysis and Experimental Results

7.1.5 Cogging Torque and Axial Force

Figure 7.11 gives the FE simulated cogging torque at 1,000rpm with the AMM stator core. Table 7.6 gives the maximum cogging torque and average axial force values. The cogging torque was approximately 6 times smaller with 10P design as compared to 4P design. This is due to the reduced interaction between the rotor and the stator slots in the fractional-slots design. In addition, the axial force is also smaller (about 20%) in the 10P design due to smaller air gap flux density.

![Figure 7.11. FE simulated cogging torque of AMM machines (4P, 10P, sintered and bonded) at 1,000rpm.](image)

As expected, the bonded magnets with lower remanent flux density resulted in a 3.5 times lower cogging torque magnitude (see Figure 7.11(b)) and 2.7 times lower axial force magnitude (see Table 7.6). Increasing the number of turns had no effect on the cogging torque or axial force. On the other hand, slightly lower values were estimated in the simulation for the sintered magnet design at 3mm airgap. This is due to the fact that the magnet is further away from the stator teeth which reduces the magnetic interaction effect.

Similar shapes of cogging torque waveforms and magnitudes were observed for the AMM, SI and SMC machines with identical physical stator dimensions. The most significant difference among the materials is the axial force magnitude as shown in Table 7.6 which depends on the airgap flux density and magnetic flux path. As the stator teeth, magnet shape and magnet remanent flux are similar, the only variable that could cause the difference in average axial force among the materials is the airgap flux...
density distribution due to the different permeability. The discrepancies are also magnified as the force is proportional to the square of the airgap flux density. Since SMC has lower permeability than AMM and SI, SMC showed the lowest axial force.

Table 7.6. FE cogging torque and axial force of AMM machines (4P, 10P, sintered and bonded), SI and SMC machines (10P, sintered).

<table>
<thead>
<tr>
<th>Material</th>
<th>Cogging Torque</th>
<th>Axial Force</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Max (Nm)</td>
<td>Average (N)</td>
</tr>
<tr>
<td>AMM</td>
<td></td>
<td></td>
</tr>
<tr>
<td>4P, sintered</td>
<td>2.18</td>
<td>1575</td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.22</td>
<td>1399</td>
</tr>
<tr>
<td>- 3mm airgap</td>
<td>0.042</td>
<td>555</td>
</tr>
<tr>
<td>10P, bonded</td>
<td>0.062</td>
<td>575</td>
</tr>
<tr>
<td>SI</td>
<td></td>
<td></td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.23</td>
<td>1413</td>
</tr>
<tr>
<td>SMC</td>
<td></td>
<td></td>
</tr>
<tr>
<td>10P, sintered</td>
<td>0.21</td>
<td>1355</td>
</tr>
</tbody>
</table>

7.1.6 Iron Loss

Figure 7.12 gives the simulated iron loss for different numbers of pole and stator and magnet materials. As given in Equation 3.7, iron loss is a function of the peak flux density and frequency. Even though the frequency in the 4P design is 2.5 times lower than the 10P design, the 4P design showed higher peak tooth and yoke flux densities and more harmonic content (see Figure 7.7). This leads to only a small difference of about 3% in the iron loss at 1,000 rpm. Nevertheless, a higher iron loss in the 10P design (up to 14%) can be observed at higher speeds (see Figure 7.12(a)).

The iron loss for the bonded magnet design is also shown in Figure 7.12(a). The lower remanent flux density of bonded magnets resulted in approximately 2 times less iron loss at 3,000 rpm, which increases with speed. Again, similar iron loss was observed for the bonded magnet configuration with higher number of turns due to similar level of flux densities. As stated before, slightly lower iron loss was observed in the simulation utilising the sintered magnet design at 3mm airgap, which is due to the lower flux densities (see Figure 7.12(a)).
In Section 3.1, it was shown that at low frequencies the grain-oriented SI material has the lowest iron loss followed by uncoated AMM and SMC materials. A similar trend was also observed in Figure 7.12(b). Nevertheless, the iron loss of SI machine increases faster than AMM machine and presents higher iron loss than AMM machine at 3,000 rpm. At 1,000 rpm, however, the iron loss in the SI stator is approximately 1.1 times lower than the AMM stator while SMC stator is about 16 times higher than to AMM stator. In addition, at 3,000 rpm, the SI stator has approximately 1.3 times and the SMC stator is about 14 times higher iron losses compared to the AMM stator. Note that, for consistency, a 17 mm slot depth AMM stator model was used to simulate the loss.

### 7.2 Construction and Testing

#### 7.2.1 AFPM Motor and Drive Specification

Figure 7.13 shows the unwound stator core on the left, and the complete motor assembly including both the windings and search coils on the right. The stator core was cut from a tape-wound toroidal AMM ribbon core using the newly developed abrasive water jet cutting technique as mentioned in Section 2.4.3. The stator core has twelve teeth, each wound with a concentrated winding. The stator coils were wound on a former and then slid into place allowing high copper packing factors to be obtained.
The rotors consist of four or ten arc-shaped, surface-mounted magnet segments with a solid mild steel back-iron as shown in Figure 7.14. The choice of four poles in the prototype required a thick rotor back-iron to avoid excessive saturation.

Figure 7.13. The stator core of the AMM-based AFPM prototype motor, without and with coils (including search coils).

Figure 7.14. Photos of four and ten pole arc-shaped magnet rotors.
Table 7.7 summarises some key parameters and dimensions of the prototype machine.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Dimensions (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator Core Material</td>
<td>2605SA1</td>
</tr>
<tr>
<td>Magnet Material</td>
<td>NdFeB</td>
</tr>
<tr>
<td>Magnet Thickness</td>
<td>3mm</td>
</tr>
<tr>
<td>Number of Turns</td>
<td>24 turn/tooth</td>
</tr>
<tr>
<td>Resistance (phase)</td>
<td>0.4Ω</td>
</tr>
<tr>
<td>DC Link Voltage</td>
<td>24V_{dc}</td>
</tr>
<tr>
<td>Outer Diameter</td>
<td>110</td>
</tr>
<tr>
<td>Inner Diameter</td>
<td>30</td>
</tr>
<tr>
<td>Base Height</td>
<td>18</td>
</tr>
<tr>
<td>Tooth Height</td>
<td>12</td>
</tr>
<tr>
<td>Shaft Diameter</td>
<td>17</td>
</tr>
<tr>
<td>Airgap Length</td>
<td>1</td>
</tr>
</tbody>
</table>

### 7.2.2 Custom-built Test Rig

Figure 7.15 shows the custom-built test setup that was specifically designed to introduce an adjustable airgap length (between 0.5mm and 8mm, with an accuracy of 0.1mm). The setup was designed to accommodate motors up to 120mm in diameter and was built with non-magnetic materials. This setup allows a number of axial-field motor configurations to be easily tested, including: single and double stator, and alternative back-iron and rotor configurations. A dc load machine was used to obtain the characteristics of the AFPM machine. The dc machine’s specifications are 36V_{dc} rated voltage, 18.3A rated current, 500W output power and rated speed of 2,500rpm. In addition, the setup was designed to achieve good stability and rigidity for high speed operation. The bearings were specially selected so that they can handle the expected high axial loading.

It should be reported here that, high speed testing was not conducted as the dc machine used for loading was only rated at 2,500rpm. In addition, a 24V_{dc}, 500W three-phase inverter was utilised for most of the tests and the 340V_{dc} three-phase inverter was only available for certain tests.
7.2.3 Motor Drive

Figure 7.16 shows the hardware implementation block diagram and the three-phase star-connected AFPM motor drive. The rotor position was measured by three Hall-effect sensors, which were positioned next to a custom-built PM sensing disk (see Figure 7.15(d)). Based on the rotor position, the appropriate stator winding phases were energised. In a star-connected three-phase winding, two phases were energised at any one time. A pulse-width modulation (PWM) technique was used to vary the inverter output voltage and hence to control the current and steady-state speed of the motor. The pulse-width modulator was made up of a sawtooth oscillator and a comparator. The commutation logic was implemented using a brushless motor controller IC. In addition, six PWM control signals were generated with the driver IC to drive the six MOSFETs in the three-phase bridge inverter.
7.3 Experimental Procedures

This section describes the experimental procedures conducted during the testing of an AFPM prototype machine based on AMM as the case study (see Section 6.3). A dc motor was used to spin the test machine in the open-circuit test and also used to act as a load in motoring mode. Sections 7.3.1 and 7.3.2 describe the procedures to obtain the open-circuit parameter characteristics of the test machine and its performance characteristics in motoring mode.

7.3.1 Parameter Measurements

The measurement of the machine parameters was performed using the setup described above. The back-EMF constant was determined from measuring the open-circuit voltage across the motor windings while the speed of the machine was varied by the dc motor. As given in Figure 7.17(a), the slope of the induced voltage versus speed curve provides the back-EMF constant of the AFPM motor.

Search coils were wound on the tooth tips and around the yoke to measure the tooth and yoke flux waveforms. The induced voltage from the search coils was passed through an RC integrator to obtain the flux waveform and scaled in terms of flux density (see Figure 6.12).

In addition, the open-circuit loss of the brushless AFPM machine was also measured using the dc machine. The dc motor was first tested by itself to determine its torque
constant and also estimate its own friction, windage and iron loss torque ($T_{\text{DC} \text{loss}}(\omega)$) as a function of speed (see Figure 7.17(b)). The dc motor was then used to drive the brushless AFPM machine under open-circuit conditions to obtain the combined dc and brushless AFPM loss torque ($T_{\text{AC} \text{loss}}(\omega)$). The difference between the combined loss curve and the dc motor loss curve gives the open-circuit loss ($T_{\text{Open} \text{loss}}(\omega)$) of the brushless AFPM motor.

A rough measurement of the peak cogging torque was conducted using the dc machine as a torque transducer due to the absence of an in-line torque transducer. The minimum dc machine armature current required to start the dc machine was measured with the dc machine both coupled to, and uncoupled from, the ac machine under test. The difference in armature current multiplied with the torque constant of the dc machine gives the peak cogging torque. A number of tests were conducted at different initial rotor positions and the measured upper and lower values were represented as horizontal lines in the cogging torque waveform (see Figure 6.16).

The inductance of the test motor was measured using a standstill test by applying a single-phase 50Hz sinusoidal ac voltage between phases A and B joined together, and phase C as shown in Figure 7.18. This connection arrangement takes into account the mutual coupling between all three phases, and the measured inductance in this configuration is equal to 1.5 times the phase inductance. The power analyser was used to measure the phase inductance. Alternatively, the reactance can be calculated and hence the inductance from the previously measured impedance and resistance. The
dc resistance of the windings was determined using a dc power supply applied to the
winding and measuring the voltage drop and current.

![Inductance measurement circuit](image)

**Figure 7.18.** Inductance measurement circuit.

### 7.3.2 Inverter Testing

The performance of the AFPM motor in six-step operation was examined using the
motor controller described in Section 7.2.1. In the experimental test set-up, the load
on the test machine was varied by changing the resistive load connected to the dc ma-
chine. The test machine output torque was then calculated utilising the dc motor as the
torque transducer (see Equation 7.1). The characteristics of the dc motor was first con-
ducted to obtain the torque constant (back-EMF constant) and dc loss torque as shown
in Figure 7.17. The three-phase input power ($P_{in}$) of the test machine was measured
using the power analyser. Alternatively $P_{in}$ can be estimated from the open-circuit loss
$P_{op}$, copper loss $P_{copper}$ and output power as given in Equation 7.2. It should be noted
that all the measured torque and efficiency shown in this thesis were estimated using
the equations.

$$T_{out} = k_{dc}I_{dc} + T_{DCloss}$$  \hspace{1cm} (7.1)

$$P_{in} = T_{out}\omega_M + P_{op} + P_{copper}$$  \hspace{1cm} (7.2)

Then, the efficiency of the test machine was calculated as:

$$Efficiency \ \eta = \frac{T_{out}\omega_M}{P_{in}}$$  \hspace{1cm} (7.3)

where $T_{out}$: output torque, $k_{dc}$: dc motor torque constant, $I_{dc}$: dc machine armature
current and $T_{DCloss}$: dc motor loss torque at the given speed $\omega_M$. 

7.4 Experimental Results of the AFPM Machine

A prototype AFPM motor was constructed based on the design described in Chapter 5. Sections 7.3.1 and 7.3.2 described the testing procedure in detail to obtain the parameter characteristics of the motor open-circuit and the performance characteristics in motoring mode. The AFPM motor was star-connected and driven by an inverter at 24V\text{dc}. As mentioned previously, the 12\,mm slot depth AMM stator designed for 12S4P machine was also utilised for the 12S10P machine configuration.

The performance of the test machine was measured as a function of operating speed with both the 24V\text{dc} and 340V\text{dc} inverters where the inverter was set to give no-load speeds of 640rpm and 2,500rpm respectively.

7.4.1 Effect of Airgap Length

The airgap length is an important parameter in axial-field PM machines as it affects both the machine’s output torque and efficiency. Section 4.4.2 showed the results for a small 32\,mm diameter machine. In this section, the characteristics of the 110\,mm diameter machine with 1\,mm and 3\,mm airgap lengths will be investigated.

Table 7.8 summarises the values of the measured back-EMF constant, cogging torque and two different no-load speeds under 1\,mm and 3\,mm airgap operations. The maximum no-load speed achieved with the 24V\text{dc} inverter was 650rpm and up to 8,500rpm using the 340V\text{dc} inverter. However, due to the speed limit of the dc load motor 2,500rpm was regarded as the high loading speed. As can be seen from the table, the measured back-EMF constant dropped by 37% and the cogging torque halved with the larger airgap length. In addition, the FE simulated axial force was also included in the table. The smaller axial force with the 3\,mm airgap would also reduce the axial loading bearing loss.

The increase in airgap length has the effect of reducing the open-circuit loss as shown in Figure 7.19(a). This is due to the reduction in the axial load bearing loss, iron loss and rotor eddy-current losses. About 40% reduction in open-circuit loss was observed at 3,000rpm. A similar pattern was observed for the no-load loss as given in Figure 7.19(b) but an approximately 40% higher loss was shown due to additional losses produced by inverter operation. The rotor eddy-current loss was investigated further using FE simulations in Section 8.3.1.
Table 7.8. Measured back-EMF constant, cogging torque and no-load test speeds for 1\textit{mm} and 3\textit{mm} airgap lengths.

<table>
<thead>
<tr>
<th>Airgap length \textit{(mm)}</th>
<th>Back-EMF constant \textit{(V/rad/s)}</th>
<th>Cogging Torque Max \textit{(Nm)}</th>
<th>No-Load Speed Low \textit{(rpm)}</th>
<th>No-Load Speed High \textit{(rpm)}</th>
<th>FE Simulated Axial Force \textit{(N)}</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.15</td>
<td>0.34</td>
<td>640</td>
<td>2,500</td>
<td>1400</td>
</tr>
<tr>
<td>3</td>
<td>0.094</td>
<td>0.19</td>
<td>640</td>
<td>2,500</td>
<td>840</td>
</tr>
</tbody>
</table>

(a) Measured open-circuit loss

(b) Measured no-load loss

**Figure 7.19.** Measured open-circuit loss and no-load loss versus speed characteristics.

Figures 7.20 and 7.21 show the open-circuit losses (estimated from curve-fitting the measured results in Figure 7.19(a)) and the copper losses obtained from the measured motor current at the range of speeds tested using the inverters. As it can be seen in the figure, the copper loss was found to be dominant at low speed and the open-circuit loss was dominant at high speed.

Figure 7.22 shows the efficiency versus output torque characteristics obtained for both airgap lengths. The figure indicates that the test machine has lower efficiency at the low speed, high torque region with an airgap length of 3\textit{mm}. However, it is able to run at higher efficiency when loaded at high speeds. In Figure 7.22(a), at low torque (<0.3Nm), the two airgaps give similar efficiency. However, as the output torque increases the efficiency at 3\textit{mm} airgap dropped much faster compared to at 1\textit{mm} airgap. This is mostly due to the fact that much higher torque is produced with an airgap of 1\textit{mm} airgap as shown in 7.23(a). On the other hand, in Figure 7.23(b) at the high loaded speed, higher torque was produced with the 3\textit{mm} airgap length over most of the speed.
range due to the machine’s lower inductance. The highest efficiency achieved at low speed is 84% with 1\text{mm} airgap length and 79% at high speed with 3\text{mm} airgap length.

![Open-Circuit Loss vs Speed](image1.png)

Figure 7.20. Measured open-circuit loss versus speed characteristics for airgap 1\text{mm} and 3\text{mm}.

![Copper Loss vs Speed](image2.png)

Figure 7.21. Measured copper loss versus speed characteristics for airgap 1\text{mm} and 3\text{mm}.

Figures 7.24 and 7.25 show the copper loss, open-circuit loss, torque and efficiency characteristics as a function of the stator mmf over different loading speeds. The test machines were operated over the mmf range of 95 to 470\text{Ampere − turn}. Figure 7.24(a) shows that the copper losses were identical for the same stator mmf. Hence, the output torque and open-circuit loss determine the efficiency. The open-circuit loss plot in Figure 7.24(b) indicates about 25% higher loss for the 1\text{mm} airgap length at low mmfs, and as the mmf increases the difference increases up to 40%. On the other hand, with
a 1 mm airgap and the low load speed, higher torque was produced compared to the 3 mm airgap with a magnitude difference of up to 45% as the mmf increases (see Figure 7.23(a)). A similar variation was observed for the high load speed.

In the efficiency plots in Figure 7.25(b), for low loaded speeds, the efficiency is significantly higher (5-10%) at 1 mm airgap. This is due to the fact that the open-circuit losses were comparable and higher output torque was generated. For the high loading speed, the efficiency is substantially higher (5-20%) for an airgap of 3 mm which is due to the much lower differences in open-circuit loss compared to the difference in output torque. The highest efficiency of 80% was obtained for the low loaded speed with an airgap of 1 mm and also for the high loaded speed with an airgap of 3 mm.
Figure 7.24. Measured copper and open-circuit loss for airgap 1mm and 3mm.

Figure 7.25. Measured output torque and efficiency for airgap 1mm and 3mm.

7.4.2 Effect of Number of Poles

Table 7.9 gives the measured back-EMF constant, cogging torque and maximum no-load speed of the 4P and 10P machines, which are consistent with the simulation results shown previously. A smaller back-EMF constant and a higher cogging torque were predicted for the 4P machine. As it can be observed in Table 7.9, compared to the 10P machine, the back-EMF of the 4P machine is 30% smaller (see Figure 7.26(a)) and the cogging torque is 10 times greater.
Table 7.9. Measured back-EMF constant, cogging torque and no-load speed for four and ten-poles machines.

<table>
<thead>
<tr>
<th>Number of Pole</th>
<th>Back-EMF Constant ((V/\text{rad/s}))</th>
<th>Cogging Torque ((\text{Min (Nm)}))</th>
<th>Cogging Torque ((\text{Max (Nm)}))</th>
<th>No-Load Speed ((\text{Max (rpm)}))</th>
</tr>
</thead>
<tbody>
<tr>
<td>4</td>
<td>0.10</td>
<td>0.33</td>
<td>2.18</td>
<td>1173</td>
</tr>
<tr>
<td>10</td>
<td>0.14</td>
<td>0.16</td>
<td>0.22</td>
<td>640</td>
</tr>
</tbody>
</table>

Figure 7.26(b) shows that comparable measured open-circuit losses were obtained for the 4P and 10P machines. Nevertheless, the simulated stator iron, magnet and rotor back-iron eddy-current losses (see Sections 7.1.6 and 8.3.1) for the 4P machine were lower than the 10P machine. Therefore, a smaller open-circuit loss was expected for the 4P machine. It was deduced that the 4P machine has higher bearing loss. In order to avoid saturation, the 4P machine back-iron thickness was doubled compared to the 10P machine and hence the weight of the rotor increased. Consequently, this increases the bearing loss due to the higher radial loading. The measured bearing loss under a radial rotor load only is shown in Figure 7.27(a) which will be analysed in details in Section 8.2.1.

![Figure 7.26](image_url)

(a) Measured phase back-EMF constant  
(b) Measured open-circuit loss

Figure 7.26. Measured phase back-EMF constant and iron loss versus speed characteristics.

Figure 7.27(b) gives the adjusted open-circuit loss which is the measured open loss (see Figure 7.26(b)) minus the measured bearing loss due to the radial loading (see Figure 7.27(a)). As can be seen in the figure, the 10P machine shows higher adjusted open-circuit loss compared to the 4P machine. The adjusted open-circuit loss consists of...
the windage loss, additional bearing loss due to the axial loading, the stator iron loss, and the rotor magnet and back-iron eddy-current losses. The windage loss is expected to be comparable as the 4P and 10P rotors have similar inner and outer diameters (see Section 8.2.2). From the simulations, the stator iron losses were comparable at low speed and slightly higher total rotor eddy-current losses were predicted for the 12S10P fractional-slots configuration (see Sections 5.9.3 and 8.3.1). Therefore, it can be concluded that the higher adjusted open-circuit loss in the 10P machine could be due to the higher rotor eddy-current losses. Nevertheless, a higher axial force was simulated for the 4P machine (see Section 7.1.5) which would increase the bearing loss due to the axial loading as discussed previously in Section 4.3.4.

This was further investigated by separating the bearing loss due to the axial loading from the measured open-circuit loss as performed in Section 4.3.4. Figure 7.28(a) shows the measured bearing loss due to radial forces only, the measured windage loss obtained using an identical size non-magnetic dummy stator core, the FE simulated stator iron, and the FE simulated rotor eddy-current losses added in a cumulative fashion. The sum is compared with the measured total open-circuit loss. A high contribution of the total loss was from the rotor magnet and back-iron eddy-current losses. Further analysis of the simulated rotor eddy-current loss is given in Section 8.3.1. The difference between the top two curves in Figure 7.28(a) is the estimated additional bearing loss due to the axial force and this is given in Figure 7.28(b). The estimated additional bearing loss of the 10P machine was also included in the figure. Even though higher axial force was simulated for the 4P machine (see Table 7.6), the estimated additional
bearing loss due to axial force of the 4P machine was up to 20% lower compared to the 10P machine. It was concluded that the larger weight in the 4P machine reduced the extra losses produced by the axial force. As a result, it is believed that the higher adjusted open-circuit loss in the 10P machine shown in Figure 7.27(b) is mainly due to the increased rotor eddy-current losses.

![Graphs showing open-circuit breakdown and bearing loss](image)

(a) Open-circuit breakdown of 4P machine  
(b) Estimated bearing loss due to axial force, 4P and 10P machines

**Figure 7.28.** Estimated open-circuit breakdown and axial bearing loss for 4P and 10P machines.

In order to examine the performance in the motoring mode, both the 4P (‘4P’ in the figures) and 10P machines (‘10P L’ in the figures) were driven by the 24\(V_{dc}\) 500W square-wave inverter. The maximum no-load speed before loading is given in Table 7.9. Figures 7.29 and 7.30 show the power loss components, output torque and efficiency plots. As the operating speeds are different, the configurations are compared in terms of the stator mmf. In addition, the 10P machine was also tested at higher speeds (‘10P H’ in the figures) using the 340\(V_{dc}\) 1kW square-wave inverter and the results are included for reference.

It can be seen in Figure 7.29(a), the copper losses are similar (about 8% difference) for both configurations due to the equal phase resistances. The open-circuit loss however, is 2 times higher in the 4P machine than the 10P machine, which is due to the higher operating speed. On the other hand, the output torque is substantially smaller in the 4P machine, which is due to the lower back-EMF constant. Overall, the 10P machine is running at higher efficiency compared to the 4P machine due to the lower open-circuit losses and the higher output torque (see Figure 7.30(a)). A performance improvement at low current was observed when the 10P machine was operated at lower speed though the peak efficiency was unchanged.
It was concluded that the lower efficiency was obtained in the 4P design that is due to its higher operating speed and hence open-circuit losses including bearing loss. This demonstrates the importance of the bearing loss. Further open-circuit loss analysis will be provided in Section 8.1.

![Copper Loss](image1.png)

(a) Measured copper loss

![Open-Circuit Loss](image2.png)

(b) Measured open-circuit loss

**Figure 7.29.** Measured copper and open-circuit losses.

![Output Torque](image3.png)

(a) Measured output torque

![Efficiency](image4.png)

(b) Measured efficiency

**Figure 7.30.** Measured output torque and efficiency.
7.4.3 Sintered and Bonded Magnets

In the FE simulation of the bonded magnet design, the number of turns were increased from 24 (sintered magnet) to 34 to achieve the same magnet flux-linkage and torque. However, only the 24 turn AMM stator was available for testing. In order to compare the performance, the airgap length of the sintered magnet design was increased to 3\textit{mm} to match the back-EMF constant of the bonded magnet design at a 1\textit{mm} airgap length (see Figure 7.31(a)). Figure 7.31(b) shows the measured no-load open-circuit losses of two types of magnets which is comparable at speeds lower than 1250\textit{rpm}. The predicted iron losses were also comparable due to the similar airgap flux densities (see Figure 7.12(a)). In addition, the windage and bearing losses are expected to be similar with identical rotor diameters and axial force (see Sections 7.1.5 and 8.2.2). However, higher rotor eddy-current losses were simulated for the sintered magnet design due to its higher conductivity and hence higher loss was expected (see Section 8.3.1). The slight increase in the total open-circuit loss at higher speeds for the sintered magnet design supports this.

![Graphs showing measured phase voltage and open-circuit loss versus speed for sintered magnet design with 3\textit{mm} airgap and bonded magnet design with 1\textit{mm} airgap.](a) Measured phase voltage (b) Measured open-circuit loss

\textbf{Figure 7.31.} Measured phase voltage and open-circuit loss versus speed for sintered magnet design with 3\textit{mm} airgap and bonded magnet design with 1\textit{mm} airgap.

As the weight of the bonded magnet rotor was higher due to the increase magnet thickness, a higher bearing loss was expected as discussed in the previous section. Figure 7.32(a) confirmed the increase of bearing loss due to radial force. Even after the bearing loss due to radial forces is subtracted from the open-circuit loss, the sintered magnet design still has a higher loss as is shown in Figure 7.32. The results are consistent with
the predictions of higher rotor magnet and back-iron eddy-current losses in the sintered magnet design. Further discussions on eddy-current loss is presented in Section 8.3.1.

The performance in motoring mode was examined and Figure 7.33(a) shows the output torque versus speed plot. The bonded magnet design was driven by a 24V_{dc} inverter and a maximum speed of 1,000rpm was reached under loaded conditions. The sintered magnet motor was driven by a 340V_{dc} inverter which was set to give a no-load speed of 650rpm during the testing period. It was found that the sintered magnet design is capable of producing higher torque above 200AT as shown in Figure 7.33(b).

![Graph](image1.png)

(a) Measured radial load bearing loss  
(b) Measured normalised loss

**Figure 7.32.** Measured bearing (radial) and normalised loss versus speed.

![Graph](image2.png)

(a) Measured output torque (speed)  
(b) Measured output torque (AT)

**Figure 7.33.** Measured output torque versus speed and mmf.
Figures 7.34 and 7.35 show the power loss components and efficiency plots. As it can be seen in Figure 7.34(a), due to the lower operating speed, the open-circuit loss was approximately 7W lower in the sintered magnet machine. In addition, 16% higher copper loss was found compared to the bonded magnet design for the same output torque as shown in Figure 7.34(b). It was expected that the sintered magnet design would perform better due to its higher torque, lower open-circuit loss and only slight increase in copper loss.

(a) Measured open-circuit loss  
(b) Measured copper loss

**Figure 7.34.** Measured open-circuit loss and efficiency versus output torque.

(a) Measured efficiency (torque)  
(b) Measured efficiency (AT)

**Figure 7.35.** Measured efficiency versus output torque and mmf.

However, as shown in Figure 7.35, the bonded magnet design was running at about 10% higher efficiency. It was deduced that the lower efficiency in the sintered magnet design was due to the lower output power (at low operating speed) combined with the
increase in the rotor eddy-current loss. The increase rotor eddy-current loss is likely due to the winding spatial and current harmonics (not measured in the open-circuit loss test) especially in the solid rotor back-iron (see Section 8.3.1). This is consistent with the simulation results showed above. The test results also showed that the eddy-current loss in the bonded magnet design was reduced significantly. This was deduced as at higher loading speeds higher eddy-current loss and open-circuit loss were expected. Nevertheless, higher efficiency was observed under these conditions with the bonded magnet design indicated that the increase in eddy-current loss was small.

### 7.4.4 Alternative Stator Materials

As discussed in the beginning of the chapter, the 17$mm$ slot depth 12S10P AMM stator was not available and a 12$mm$ slot depth 12S10P AMM stator was used in the tests. In addition, the SMC and SI stators with 17$mm$ slot depth were machined as reported in Section 5.11. However, the FE simulation results generated for both 12$mm$ and 17$mm$ slot depths AMM machines in Section 7.1.1 showed that the performance is comparable with only about 6% difference in the overall efficiency. The 10-pole AFPM SI and SMC machines were wound with the same number of turns as the AMM machine, and driven by the 24$V_{dc}$ inverter. Note that, although the grain-oriented silicon-iron laminations (SI) have better properties compared to conventional non-oriented silicon-iron, it is not commonly applied in electrical machines.

Figure 7.36 gives the induced voltage and open-circuit loss versus speed plots for the three stator material types. The results presented in the figure show that the back-EMF constants were comparable for all the designs. In addition, the bearing, windage and rotor eddy-current losses were expected to be comparable. Therefore, the differences in the measured open-circuit losses shown in Figure 7.36(b) is largely due to the stator iron loss. As predicted, the SMC design has the highest no-load open-circuit loss. The open-circuit loss of the SI machine was found slightly larger at higher speed compared to the AMM machine. This could be due to the lower stator yoke iron loss in the 12$mm$ slot depth AMM stator design (see Section 7.1.1). However, as reported previously the AMM laminations utilised to construct the stator were not insulated properly, and the high winding tension during manufacturing also increases the iron losses of the AMM stator design. It is believed that the AMM design has the potential for very low iron loss if a properly insulated AMM core is used in the design.
In order to examine the performance characteristics of the three test machines, the power loss components were obtained and the results are given in Figures 7.37 to 7.39. From the output torque plots (see Figure 7.37), the SI design showed highest output torque at a given operating speed but at the same mmf the AMM design produced approximately 0.2 Nm higher torque. Even though a higher current was drawn at the same output torque with the SMC design, the copper loss generated was comparable to the AMM design as shown in Figure 7.38. This is due to the lower winding resistance of SMC stator.

Based on the measured open-circuit loss curve given in Figure 7.36(b), the loss at different output torques corresponding to the results shown in Figure 7.37(a) was obtained as shown in Figure 7.39(a). Approximately 40% higher open-circuit loss was produced in the SMC stator design which was the main factor causing its substantially lower efficiency.

As for the SI design, the output torque (Figure 7.37(b)), the input current (Figure 7.38(b)) and the open-circuit loss characteristics (Figure 7.39(a)) are comparable to the AMM design. However, the AMM design showed higher copper loss (see Figure 7.38(a)) due to its higher stator winding resistance associated with the shallower stator slots. As a result, the SI design had similar efficiency as the AMM design at low torque (< 0.5 Nm). However, as the torque increases, the SI design is able to operate at slightly higher efficiency (see Figure 7.39(b)). This is consistent with the simulation results where both SI and AMM stators had similar performance characteristics, except the iron and copper losses characteristics which is due to the differences in the stator winding resistance.
Figure 7.37. Measured torque versus speed and mmf characteristics.

(a) Measured output torque (speed)  
(b) Measured output torque (AT)

Figure 7.38. Measured copper loss and input current versus output torque characteristics.

(a) Measured copper loss  
(b) Measured input current

Figure 7.39. Measured open-circuit loss and efficiency versus output torque characteristics.

(a) Measured open-circuit loss  
(b) Measured efficiency
Chapter 7  110mm Machine 3D FE Analysis and Experimental Results

7.5 Conclusions

In this chapter, a 110mm diameter AFPM motor utilising cut AMM was designed, constructed and studied. The machine’s parameters (i.e. flux densities, induced back-EMF, inductances, cogging torque, axial force and iron loss) were predicted using a 3D FEA tool and tested using a custom-built test setup. In addition, the performance characteristics of the motor were examined and presented. This includes comparisons of the effects of different airgap lengths, number of poles, magnet and stator magnetic materials.

It was demonstrated that the analytical calculation method presented in Chapter 5 showed reasonable accuracy, and good agreement was obtained in the 3D FEA modelling approach compared to the measured results. The experimental results indicated that copper loss is the dominant loss at low speeds but as the speed increases the open-circuit loss, in particular the stator iron and rotor eddy-current losses become the dominant factor.

In addition, it was found that increasing the airgap length reduces the air gap flux that is produced by the rotor magnets. Therefore, it was observed that the back-EMF, cogging torque, stator iron losses and axial force were also reduced at larger airgap lengths. The bearing loss is smaller due to the lower axial force. However, for the same output torque, the smaller back-EMF would produce higher copper loss due to increased winding current. Furthermore, increasing the air gap reduces the airgap flux that is due to the stator currents and hence reduces the inductance. As a result, for the same speed, particularly at high operating speeds, higher current is required, which causes higher copper losses. Therefore, it can be concluded that there is a tradeoff in the overall performance of the machine at different airgap lengths.

Based on the FE simulation and the test results, the 10-pole design was found to have similar stator iron loss with the 4-pole design although it has a fundamental frequency of 2.5 times higher. This is due to the fact that the 4-pole design has higher flux density magnitude and harmonic content. In addition, comparing the test results based on the 4 pole and 10 pole rotors of different weights, a high bearing loss was found in the 4-pole design which affects the overall efficiency of the machines. Even though the magnet axial force increases the axial loading of the bearing and hence bearing loss, it was not expected that the bearing loss was mainly affected by the weight of the rotor. As the rotor weight increases the bearing loss increases, but becomes less sensitive to axial loading.
High rotor magnet and back-iron eddy-current losses were calculated for the fractional-slot configuration. The use of bonded magnets result in lower magnet eddy-current loss. However this was offset by the increased rotor weight due to the thicker magnets which increased the bearing loss. Overall the bonded magnet design (with 1\text{mm} airgap) had about 10\% higher measured efficiency than the sintered magnet design (with 3\text{mm} airgap). It should also be noted that the thicker bonded magnet design also increases the motor’s axial length compared to the sintered magnet design.

It was observed that the AMM prototype machine did not present higher efficiency at low speed, but it was still able to achieve comparable efficiency reference to the similar types of machine as given in Table 2.7. This was expected in a design that has non-optimised construction of the AMM stator which produces higher iron loss due to non-insulated AMM strip. It is believed that an optimised AMM machine can offer much higher efficiency at higher speeds where iron loss is the dominant loss. This was deduced from the tests utilised with the coated AMM ribbon ring core (see Section 3.1).
IN this chapter, the identified high open-circuit loss with the 32\textit{mm} prototype was examined further using the 110\textit{mm} machine. Specific bearings were designed to handle the expected axial loading, which were utilised in the construction of the test rig. Additional slotted and dummy cores including a double-sided rotor were also constructed for the analysis of the losses. To analyse and separate the open-circuit loss components, both experimental and 3D FE modelling were conducted and presented in this chapter.
8.1 Open-Circuit Loss Analysis

As reported, in Sections 4.3.4 and 7.4, high open-circuit loss was measured in the single-sided machine configuration. The open-circuit loss of the machine consists of the stator iron, bearing, windage and magnet and rotor back-iron eddy-current losses. In addition, the bearing loss has two components: due to the radial load (rotor weight) and an additional component due to the axial loading.

In this section, a double-sided stator configuration was utilised to balance the axial force on the rotor and to estimate the axial component of the bearing loss. In addition, an identical size non-magnetic dummy stator core was included in the configuration to estimate the windage loss. Furthermore, the stator core iron loss and the magnet and rotor back-iron eddy-current losses were simulated by FE method.

The open-circuit loss in this study was obtained by measuring the input torque required to drive the PM rotor under open-circuit conditions. The input torque was estimated by measuring the input armature current of the dc dynamometer motor both without load and when driving the PM rotor. As explained in section 7.3.1, the input torque is the difference between the two armature current readings multiplied by the DC machine’s torque constant.

The following sections report the results of the investigations to separate the above mentioned loss components.

8.2 Bearing and Windage Losses

8.2.1 Bearing Loss

Radial Load

As stated above, the bearing loss caused by the radial load is due to the weight of the rotor including magnets. This loss can be obtained by measuring the input torque and hence power required to spin the rotor in free air (with no stator present). It should be noted that, such measurement method includes some windage loss as well.

The measured radial bearing loss \( P_{\text{Radial loss}}(\omega) \) of different rotors are plotted in Figure 8.1 as a function of speed. As given earlier in Equation 5.77, the radial load bearing loss is linearly proportional to both the rotor weight and the rotor speed. This was also observed in Figure 8.1 where the heaviest rotor (4P) test presented highest loss.
Chapter 8  110mm Machine Open-Circuit Loss Analysis

![Graph showing bearing loss due to radial load of four rotors of different weights.](image)

**Figure 8.1.** Bearing loss due to radial load of four rotors of different weights. Single-sided 4 pole (sintered, 2.28 kg), single-sided 10 pole (bonded, 1.95 kg), double-sided 10 pole (sintered, 1.77 kg) and single-sided 10 pole (sintered, 1.57 kg).

**Axial Load**

To measure the additional bearing loss component caused by the axial loading, a double-sided rotor was constructed (using the same back-iron thickness as in the single-sided rotor configuration) and a second stator core was utilised on the other side of the double-sided rotor to counter-balance the axial force, which is illustrated in Figure 8.2. During the tests, a AMM stator (slot depth of 12 mm) and a SI stator (slot depth of 17 mm) were used.

![Diagram showing single-sided and double-sided rotor configurations with stators.](image)

**Figure 8.2.** a) single-sided and b) double-sided, axial-field motor configurations with SI, and SI and AMM stators.
Chapter 8  110\text{mm} Machine Open-Circuit Loss Analysis

The procedure used in the tests involves the following steps.

- Obtain the open-circuit loss as a function of speed for the single-sided AMM stator with a 1\text{mm} airgap.

- Subtract the radial bearing loss measured with the rotor rotating in free air to obtain $P_{AMMOploss}(\omega)$.

- Repeat the above procedure for the single-sided SI stator with a 1\text{mm} airgap to obtain $P_{SIOploss}(\omega)$, and for double-sided configuration using AMM and SI stators each with 1\text{mm} airgap to give $P_{AMMSIOploss}(\omega)$.

Assuming that the stator iron and rotor eddy-current losses in the double-sided configuration is twice of one single sided configuration, the additional bearing loss component $P_{Axialloss}(\omega)$ that is due to the axial force is estimated by

$$P_{Axialloss}(\omega) = 0.5[(P_{AMMOploss}(\omega) + P_{SIOploss}(\omega)) - P_{AMMSIOploss}(\omega)] \quad (8.1)$$

Table 8.1 gives the FE simulated stator iron and rotor eddy-current losses for the machine configurations mentioned above. The results demonstrated that the above assumption is acceptable.

**Table 8.1.** FE simulated single and double-sided configurations stator iron and eddy-current losses at 1,000rpm.

<table>
<thead>
<tr>
<th>Losses</th>
<th>Single-sided AMM stator $P_{AMMOploss}$ (W)</th>
<th>Single-sided SI stator $P_{SIOploss}$ (W)</th>
<th>Double-sided AMM &amp; SI stators $P_{AMMSIOploss}$ (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Iron loss</td>
<td>1.02</td>
<td>1.04</td>
<td>2.11</td>
</tr>
<tr>
<td>Eddy-current loss</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>- magnet</td>
<td>6.41</td>
<td>6.41</td>
<td>12.9</td>
</tr>
<tr>
<td>- back-iron</td>
<td>4.5</td>
<td>4.5</td>
<td>8.3</td>
</tr>
</tbody>
</table>
Figure 8.3 shows the measured loss versus speed results for the sum of the two single-sided configurations \(P_{AMMOploss} + P_{SIOploss}\) and also for the double-sided configuration \(P_{AMMSIOploss}\). The difference between the two plots is the bearing loss due to axial loading as given in Figure 8.3(b). The result showed that the additional bearing loss due to axial loading is approximately 5 times lower than the bearing loss due to radial force (see Figure 8.1). The axial force related to bearing loss estimated using the method described in Section 7.4.2 is also given in Figure 8.3(b) which shows a maximum discrepancy of about 15%.

![Figure 8.3](image)

(a) Open-circuit losses  
(b) Axial load bearing loss

Figure 8.3. a) Open-circuit losses of single and double-sided configurations after the bearing loss due to radial loads have been subtracted, b) bearing loss component due to axial force.

**Analytical Comparison of Bearing Losses in the 12S10P Machine**

The analytically calculated bearing loss for the radial load using Equation 5.77 is shown in Figure 8.4(a). The maximum coefficient of bearing friction \(k_{fb}\) value of the equation for small ball bearings was \(3m^2/s^2\) and the bearing loss was about 1.5 times smaller than the measured value. In order to match the measured bearing loss due to radial loads, a value of \(k_{fb}\) of 4.7 was required. Figure 8.4(b) and Table 8.2 summarise the required \(k_{fb}\) values to match the measured power loss results of the rotors with different weights. As it can be seen in the figure, the \(k_{fb}\) value increases with the rotor weight.
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110\(\text{mm}\) Machine Open-Circuit Loss Analysis

(a) Bearing loss  

(b) Measured coefficient of bearing friction \(k_{fb}\)

![Graph](image)

Figure 8.4. Bearing loss comparison.

Table 8.2. Coefficient of bearing friction \(k_{fb}\) versus weight values from Figure 8.4(b).

<table>
<thead>
<tr>
<th>Rotor Weight</th>
<th>(k_{fb})</th>
</tr>
</thead>
<tbody>
<tr>
<td>(kg)</td>
<td>((m^2/s^2))</td>
</tr>
<tr>
<td>1.57 (10P, sintered, single)</td>
<td>4.7</td>
</tr>
<tr>
<td>1.77 (10P, sintered, double)</td>
<td>5.0</td>
</tr>
<tr>
<td>1.95 (10P, bonded, single)</td>
<td>5.5</td>
</tr>
<tr>
<td>2.28 (4P, sintered, single)</td>
<td>5.8</td>
</tr>
</tbody>
</table>

8.2.2 Windage Loss

A non-magnetic dummy stator with a 1\(\text{mm}\) airgap was used to measure the open-circuit loss. It was assumed that the windage loss of the rotor rotating in free air is zero. Then, it was found that the increased loss with the dummy stator (compared to \(P_{\text{Radial loss}}\), see Section 8.2.1) gives the windage loss (\(P_{\text{Wind loss}}\)) at the fixed airgap length.

Figure 8.5(a) gives the measured windage loss as a function of speed at an airgap of 1\(\text{mm}\). In addition, the analytically calculated value from Equation 5.74 is also included in the figure, which is much lower than the measured value. The windage loss was found about 15 times lower than the bearing loss as shown in Figure 8.5(b).
8.3 Iron and Eddy-Current Losses

During the experiments, the rotor back-iron temperature of the machines increased from a room temperature of 24°C up to 55°C at 8,500 rpm (no-load). Hence, it was concluded that there were high eddy-current loss in the magnets and the back-iron.

The open-circuit loss measured in the single-sided configuration after subtracting the estimated bearing and windage losses from previous section is given in Figure 8.6(a). This represents the measured stator iron and rotor magnet and back-iron eddy-current losses. Also included in the figure are the FE simulated stator iron loss and the sum of this loss plus the FE simulated rotor eddy-current losses. It was found that the calculated eddy-current loss in the magnets and back-iron is about 10 times higher than stator iron loss, and there is a good agreement between the calculated and the measured total stator iron and rotor eddy-current losses.

Figure 8.6(b) gives all the open-circuit loss components added cumulatively. The major components of the loss are the bearing loss followed by the rotor eddy-current loss. The total loss was within 5% of the measured value. In addition, a similar loss breakdown was conducted on the results measured in the SI and SMC machines as shown in Figure 8.7. The calculated losses were within 12% for SI machine and within 5% for SMC machine compared to the measured values.
Figure 8.6. a) FE simulated and measured stator iron and rotor eddy-current losses of the AMM stator, b) open-circuit losses breakdown for the AMM stator.
Chapter 8  110mm Machine Open-Circuit Loss Analysis

Figure 8.7. FE and measured stator iron and rotor eddy-current losses: a) SI and b) SMC materials.
8.3.1 Eddy-Current Loss Analysis

As discussed in Section 5.9.3, eddy-current losses in the rotor back-iron and surface-mounted permanent magnets for fractional-slot designs can be very significant. The large amplitude spacial winding sub-harmonics, the stator current time harmonics and the space harmonics due to slotting effects can induce large rotor eddy-current losses especially at high speed operation.

3D FEA was utilised to simulate the eddy-current losses induced in the magnets and back-iron. In order to evaluate the slotting effects, the eddy-current loss generated during open-circuit conditions was simulated. Table 8.3 gives the simulated average eddy-current losses in the sintered magnets and the rotor back-iron for non-slotted and slotted cores with various airgap lengths. As it can be seen in the table, in the slotted core, the solid back-iron loss was comparable to the total magnet loss. However, the losses dropped substantially particularly in the rotor magnets, as the airgap length was increased from 1\( mm \) to 3\( mm \) and then to 8\( mm \).

<table>
<thead>
<tr>
<th>Slotted, airgap length (( mm ))</th>
<th>Eddy-current Loss</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Magnet (W)</td>
<td>Solid Back-Iron (W)</td>
</tr>
<tr>
<td>Non-slotted</td>
<td>0.049</td>
<td>0.41</td>
</tr>
<tr>
<td>Slotted</td>
<td>3.5</td>
<td>3.34</td>
</tr>
<tr>
<td>1</td>
<td>3.5</td>
<td>3.34</td>
</tr>
<tr>
<td>3</td>
<td>0.88</td>
<td>2.06</td>
</tr>
<tr>
<td>8</td>
<td>0.53</td>
<td>1.80</td>
</tr>
</tbody>
</table>

In the non-slotted core, the magnet losses were about 71 times and the back-iron loss was about 8 times smaller than the slotted core. The losses in the non-slotted core could be due to the simulation errors as zero loss was expected in the non-slotted design at open-circuit conditions. Further work is required to explain the non-zero loss which is left for future work.
Chapter 8 110mm Machine Open-Circuit Loss Analysis

No significant changes were observed in the rotor eddy-current loss operating under different stator materials. In addition, as it was shown previously in Section 7.1.1 that changes in the stator slot depth have no significant effect on the rotor eddy-current loss.

In the following sections, the rotor eddy-current loss was examined under various conditions.

Four-Pole Machine versus Ten-pole Machine

Figure 8.8(a) gives the magnet and the back-iron eddy-current losses, and Figure 8.8(b) illustrates the total rotor loss as a function of speed for the four-pole (4P) and ten-pole (10P) machines under open-circuit conditions (O/C). As expected the results show that the eddy-current loss is proportional to the frequency squared. The magnet loss is about 1.2 times higher in the 4P machine. On the other hand, the eddy-current loss in the back-iron was about 1.2 times lower in the 4P machine. As a result, the total eddy-current loss in the 10P machine is only about 5% higher compared to the 4P machine at 3,000rpm as shown in Figure 8.8(b).

![Graph](image)

(a) FE total magnet and back-iron loss (O/C)

![Graph](image)

(b) FE total loss (O/C)

Figure 8.8. FE simulated total magnet and rotor back-iron eddy-current losses and total rotor eddy-current losses as a function of speed for the 4P and 10P designs under open-circuit conditions.
In order to evaluate the rotor eddy-current loss due to the winding spatial and stator current time harmonics, the losses were calculated under loaded condition (L) as given in Figure 8.9. As it can be seen in the figure, the average values of the losses under both open-circuit and loaded conditions are comparable.

![Graph showing simulated total magnet and rotor back-iron eddy-current losses and total rotor eddy-current losses as a function of speed for the 4P and 10P designs under open-circuit and loaded conditions.](image)

**Figure 8.9.** FE simulated total magnet and rotor back-iron eddy-current losses and total rotor eddy-current losses as a function of speed for the 4P and 10P designs under open-circuit and loaded conditions.

**Sintered versus Bonded Magnets and Solid versus Laminated Back-Iron Designs**

Table 8.4 gives the FE calculated magnet, back-iron and total eddy-current losses of sintered magnet machines with 1mm and 3mm airgaps, and a bonded magnet machine with a 1mm airgap under open-circuit conditions.

<table>
<thead>
<tr>
<th>Configurations</th>
<th>Magnet (W)</th>
<th>Back-Iron (W)</th>
<th>Total (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sintered, 1mm airgap, solid</td>
<td>30.9</td>
<td>24.9</td>
<td>55.8</td>
</tr>
<tr>
<td>Sintered, 3mm airgap, solid</td>
<td>7.64</td>
<td>5.49</td>
<td>13.1</td>
</tr>
<tr>
<td>Bonded, 1mm airgap, solid</td>
<td>1.52</td>
<td>3.99</td>
<td>5.51</td>
</tr>
<tr>
<td>Sintered, 1mm airgap, SI Lam</td>
<td>25.2</td>
<td>6.15</td>
<td>31.4</td>
</tr>
<tr>
<td>Sintered, 1mm airgap, AMM Lam</td>
<td>25.2</td>
<td>3.38</td>
<td>28.6</td>
</tr>
</tbody>
</table>
For the same airgap, the sintered magnet design showed about 20 times more magnet losses compared to the bonded magnet design, and displayed about 6 times more back-iron loss. On the other hand, when the airgap length of the sintered design is increased to 3\,mm to match the induced voltage of bonded design the losses were reduced. Nevertheless, the losses in the sintered design at 3\,mm air gap operation are 5 times higher in the magnet and 1.4 times higher in the back-iron compared to the bonded design at 3,000\,rpm.

In order to reduce the eddy-current losses, two common methods can be applied: segmenting the magnets and utilising laminations for the rotor back-iron. The latter was investigated by replacing the back-iron with the toroidally wound thin core. The laminated back-iron (8\,mm thick) was first set to use the silicon iron (mild steel) characteristic (SI Lam), and then an AMM characteristic (AMM Lam) (see Table 8.4).

For the back-iron loss, a 4 times reduction was obtained with the laminated SI and a 7.4 times reduction as obtained with the laminated AMM. A significant decrease in the average magnet loss (about 22.5\%) was observed between the solid and laminated back-iron. However, there is no significant difference in average magnet loss between the laminated SI and AMM.
8.4 Loss Component Comparison

Figure 8.10 gives the calculated loss components of the total open-circuit loss for the six AMM, SI and SMC machine configurations at 3,000 rpm and at 10,000 rpm operating speeds. The machine configurations are 10 pole AMM, SI and SMC materials with sintered magnets at 1 mm airgap (10P, AMM, SI and SMC), 4 pole AMM material with sintered magnets at 1 mm airgap (4P, AMM), 10 pole AMM material with sintered magnets at 3 mm airgap (10P, AMM, 3 mm) and 10 pole AMM material with bonded magnets at 1 mm airgap (10P, AMM, Bonded). Figure 8.11 shows the same results in Figure 8.10 with each bar graph normalised to 100%. Note that, the open-circuit losses in the figures were measured up to 3,000 rpm that is limited by the maximum speed of the loading dc machine, and the loss components were extrapolated up to 10,000 rpm.

Based on the analytical equations of the bearing and windage losses given in Section 5.9.4, the losses were extrapolated by curve fitting with a polynomial of order 3. In addition, a polynomial of order 2 was fitted to the stator iron and the rotor eddy-current losses.

At 3,000 rpm, the two dominant losses are the bearing and rotor eddy-current losses, which were roughly equal for all the configurations, except for the case at 3 mm airgap (10P, AMM, 3 mm) and the bonded magnet design (10P, AMM, Bonded). As expected, the bonded magnet has very small rotor eddy-current loss due to its high resistivity. However, similar open-circuit loss was obtained for the sintered magnet design at 3 mm airgap and bonded magnet design at 1 mm airgap, even though these two cases have similar flux density and hence similar axial loading on the bearings. This is due to the increased radial loading on the bearings causing higher loss as the bonded magnet design required thicker bonded magnets. As it can be observed in the figures, the windage loss is the lowest loss component in all machine configurations. In addition, the stator iron loss is small for both AMM and SI materials except for SMC material. As a result, the SMC has the highest total loss while the AMM with bonded magnet rotor has the lowest total loss.

As speed increases, the iron and the eddy-current losses which are proportional to the frequency squared are expected to increase rapidly. Comparing the bar graphs at 10,000 rpm, it can be seen that the rotor eddy-current loss increases to about 2 to 3 times of bearing loss and hence become the dominant loss, while the windage loss remains the lowest. Although the stator iron loss still remains small (except for the SMC), it is at a larger fraction of the total loss. In addition, the dominance of eddy-current loss
increases the difference between the total loss of bonded magnet design and the other designs.

![Graph](image)

(a) 3,000rpm measured

![Graph](image)

(b) 10,000rpm prediction

Figure 8.10. Open-circuit loss comparisons for various configurations at 3,000rpm and 10,000rpm.
Figure 8.11. Open-circuit loss comparisons for various configurations at 3,000rpm and 10,000rpm.
8.4.1 Efficiency Contour Plots

In this section, the calculated efficiency contour plots of various machine configurations were compared with the 12S10P AMM stator of 12mm slot depth (SD) that accommodated sintered magnets and operated at 1mm airgap length as the baseline design. The plots shown in Figure 8.12 were obtained using the same method described in Section 4.4. Note that, the experimental operating points of the machines are also shown on the contour plots as symbols where available. Figure 8.13 is given to illustrate the same information in colour plots. It can be concluded that in general, the estimated efficiency characteristics are about 1 to 5% higher than the measured results due to the smaller copper loss prediction as shown previously in Section 4.4. Despite this, the plots provide a means for comparing the relative performance of different designs.

Figures 8.12(a) and 8.12(b) show the calculated efficiency contour plots for the 10P sintered (3mm airgap) and bonded (1mm airgap) designs. As shown, both designs have comparable performance and a peak efficiency up to 80% at 3,000rpm. The efficiency contours are also comparable at low speed and low torque region due to similar open-circuit loss caused by the increased rotor weight (double bonded magnet thickness) and hence the bearing loss. Nevertheless, at higher speed the bonded magnet design offers higher efficiency due to smaller rotor eddy-current loss compared to sintered magnet design.

The 10P sintered design with 1mm airgap has a 2% higher efficiency than with 3mm airgap operation over most of the operating range (see Figures 8.12(a) and 8.12(c)). At higher speed and torque (>0.75Nm) operating points, higher efficiency is predicted at 3mm airgap due to low open-circuit losses.

Comparing the plots in Figures 8.12(c) and 8.12(d), it can be concluded that the 10P design has generally 2 to 5% higher efficiency over much of the operating region. In addition, the 10P design shows better performance at high torque range than the 4P design.

Figures 8.12(c), 8.12(e) and 8.12(f) show the efficiency plots of AMM, SI and SMC machines with sintered magnets and at the same airgap of 1mm. The AMM stator has slot depth of 12mm while the SI and SMC stators have slot depth of 17mm. Based on the efficiency plots, the predicted efficiencies at 1,000rpm and 1.5Nm were tabulated in Table 8.5. The results show that the grain-oriented SI machine had the highest efficiency followed by the AMM and SMC machines, which is consistent with the experimental
Figure 8.12. Efficiency contour plots. The baseline machine (c) is an AMM stator with 12mm slot depth, a 1mm air gap and a sintered 10 pole rotor. The other machine configurations indicate the diversion from this baseline design.
results (see Figure 7.39(b)). Higher efficiencies were predicted at higher speed for all machines which are consistent with the above conclusions.

![Efficiency contour colour plots](image)

**Figure 8.13.** Efficiency contour colour plots based on Figure 8.12.
Table 8.5. Efficiency values of different materials at 1,000 rpm and 1.5 Nm, which are obtained from efficiency contour plots for 1 mm airgap length.

<table>
<thead>
<tr>
<th>Material</th>
<th>1 mm (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>AMM</td>
<td>76</td>
</tr>
<tr>
<td>SI</td>
<td>78</td>
</tr>
<tr>
<td>SMC</td>
<td>74</td>
</tr>
</tbody>
</table>

Figures 8.14(a) gives the efficiency contour plots of the baseline machine under 90% reduction in the bearing and rotor eddy-current losses in the previous plot. For the bonded magnet design, the efficiency contour plots under 90% reduction of the bearing loss in previous plot is given in Figure 8.14(b). As can be seen from the figures, higher efficiency is predicted at higher speed and torque operating points with the loss reduction. In addition, the maximum efficiency increased by about 10% for the sintered magnet design and about 8% for the bonded magnet design.

![Figure 8.14(a)](image1.png)  
(a) 10P baseline design

![Figure 8.14(b)](image2.png)  
(b) Bonded, 1 mm airgap

Figure 8.14. Efficiency contour plots of the baseline machine and bonded magnet design under 90% reduction in bearing and rotor eddy-current losses.
8.5 Conclusions

The high open-circuit loss observed in the single-sided AFPM machine configuration was analysed in detail in this chapter. There are four main components in the open-circuit loss: the bearing, windage, stator iron and rotor eddy-current losses. In this study, the bearing and windage losses were separated experimentally while the stator iron and rotor magnet and back-iron eddy-current losses were calculated using 3D FE analysis. In addition, the bearing loss was further separated into the loss caused by the radial and axial loads, which were tested using a double-sided stator configuration with reduced axial force. It can be concluded that the sum of the individual losses match the measured open-circuit loss within 10% for all types of stator materials.

The following summarises the key findings of the rotor eddy-current loss analysis.

- The magnet and rotor back-iron eddy-current losses increase with the presence of slots and the rotor speed, but decrease as the airgap length increases.
- The losses under open-circuit and load conditions are comparable.
- The eddy-current loss in the rotor back-iron is generally an order of magnitude higher compared to the magnet.
- The total loss of the 10-pole design is only about 5% higher than the 4-pole design at 3,000 rpm.
- Compared to the bonded design at 3,000 rpm, the loss in the sintered magnets is about 5 times higher in the magnet and 1.4 times higher in back-iron higher.
- A significant decrease in the average magnet loss (about 22.5%) was observed with the laminated back-iron at 3,000 rpm. As for the back-iron loss, 4 times reduction with solid mild steel characteristic and 7.4 times reduction with the AMM characteristic were observed.

In addition, the loss components of the total open-circuit loss for various machine configurations were compared. In general, at 3,000 rpm, the main loss contribution is due to the bearing loss followed by the rotor eddy-current, stator iron and windage losses. Although the bearing loss is the dominant loss below 10,000 rpm, above this speed the rotor eddy-current loss exceeds the bearing loss. On the other hand, the rotor eddy-current loss of the bonded magnet design was about 10 times smaller than in the sintered magnet design. Therefore, the bearing loss remain as the dominant loss in this
case. The overall conclusion is that above 10,000rpm, the main loss contributions are 
the rotor eddy-current loss followed by bearing, stator iron and windage losses.

The calculated efficiency contour plots of the machine configurations were also com-
pared with the baseline design in this chapter to examine their relative performance 
over a wide range of torque and speed operating points. It was predicted that 90% 
reduction in the bearing and rotor eddy-current losses would improve the efficiency of 
the machine up to 10%.
Chapter 9

Conclusions and Future Work

9.1 Conclusions and Contributions

This research work has focused on utilising the capabilities of the improved cutting process for Amorphous Magnetic Material (AMM) ribbon developed by the industrial partner to design and investigate slotted AMM surface permanent magnet (PM) concentrated-winding axial-field machines. The key results achieved are summarised in Section 9.1.1. The major contributions resulting from this research work are given in Section 9.1.2.

9.1.1 Key Results of AMM AFPM Machines

The principal aim of this work was to investigate an innovative axial-field PM (AFPM) machine design based on the improved AMM cutting process. The use of AMM in the machine design was proposed to take advantage of its low iron loss characteristics specifically at higher operating frequencies.

The first stage of the research was to compare the iron losses at various frequencies for selected conventional and emerging magnetic materials. A comprehensive series of iron loss measurements were conducted using non-slotted ring core samples. The test results show that at 50Hz, the coated AMM has the lowest iron loss density followed by grain-oriented silicon iron (SI), uncoated AMM, non-oriented SI and soft magnetic composite (SMC).
The measured iron loss data for uncoated AMM was utilised in the 3D finite element modelling. In this section of study, the measured and simulated results for the non-slotted core matched closely, which provided high confidence in the modelling approach. Two compensation methods were implemented to estimate the iron loss density of slotted stator cores based on the results of the circumferential flux test. The first method was based on the average flux density in the slotted core and the total core weight. The second method used the peak yoke flux density in the core and a reduced effective core weight that was determined by FE analysis to match the iron loss densities between the non-slotted and slotted cores. It was found that the slotted stator core has 30% more iron loss density than the non-slotted stator core which is likely to be caused during the cutting process.

An analytical model was derived to investigate the ratio of the intralaminar (conventional) and interlaminar (due to lack of insulation between laminations) eddy-current losses. It was found that the calculated ratio is proportional to the interlaminar conductivity and the core dimensions. Nevertheless, the calculated interlaminar eddy-current loss accounted for only about 25% of the measured iron losses at high frequencies.

A 3D FE model was utilised to simulate the AFPM machine and to examine the machine behaviour. In order to minimise the computation time, a mesh size selection and optimisation process was described. The simulated parameters were then examined and compared with the experimental results. Overall, there was a good agreement between them and so there is strong confidence in the model representing the characteristics of the designed AFPM machine.

Although 3D FEA offers accurate performance prediction of the machine being modelled, it also requires a long computation time, which is not favorable in the initial stage of the machine design process and hence is more useful as a validation tool. Therefore, one of the objectives of this research was to develop analytical design procedures allowing rapid determination of the key design variables and analysis of the basic characteristics of the machine. This was used to take into consideration the limitations of the laboratory test rig, the available magnet type, the AMM cutting technique and the AMM saturation flux density. The design procedure proposed includes initial design guidelines to determine the core size, and to select the slot and pole combination, magnet thickness, slot depth, slot width, stator yoke thickness and number of turns in the machine. It was demonstrated that the analytical approach provides a rough initial design which could then be fine tuned and optimised in 3D FE analysis.
In general, fewer slots in the AMM core require less cutting time and hence reduces the cost. Nevertheless, as discussed in Chapter 5 a machine with better performance would generally require a high pole number which leads to higher number of slots. Therefore, there is a trade-off between the performance and cost saving depends on the requirements and application. In order to optimise the design based on shear stress and iron loss analysis, the optimum values of inner/outer diameter ratio, slot depth and width (magnetic versus electric loading) were also investigated. This was conducted for sizing purposes and to examine possible options for reducing the cutting cost in terms of shorted radial stack length and slot size. The relationship between the stator outer diameter and the length of the airgap and neodymium magnet thickness was also discussed. The analytical methods were implemented in Matlab, and magnet demagnetisation issues were investigated.

Slotted AMM stators were implemented successfully in small and medium AFPM motors. Due to available resources, the prototype had to be built using uncoated AMM ribbon which resulted in substantially higher iron losses. The prototype machine was run in a custom-built test rig that was used to conduct an extensive series of experiments and to investigate and demonstrate the characteristics of the AMM prototype machine. This included comparisons of the AMM machine with identical size SI and SMC prototypes. Overall, a good correspondence between the calculated and test results was achieved.

The 110\(mm\) diameter AMM prototype achieved 80\%\ efficiency which is comparable to the value obtained from the similar types of machine reported in literature. The small 32\(mm\) diameter AMM prototype displayed a maximum efficiency of 70\%. The experimental results of the prototype also highlighted the significance of the open-circuit losses which includes the bearing, windage, rotor eddy-current and stator iron losses on the overall machine performance. The identified components of the open-circuit loss were separated and analysed both experimentally and using FE simulations.

In the open-circuit loss analysis, it was demonstrated that, at 3,000\(rpm\), the two main loss contributions are the bearing loss and the rotor eddy-current loss which are nearly equal. The rotor eddy-current loss increases rapidly with speed, and at 10,000\(rpm\), it becomes the dominant loss followed by the bearing, stator iron and windage losses. It was shown that the eddy-current losses can be reduced by using bonded magnets and a laminated back-iron or by increasing the airgap with the sintered magnets. The efficiency of the alternative machine designs were predicted based on the measured
open-circuit loss and the measured machine parameters and were shown in contour plots. From the contour plots, up to 10% increased in efficiency was predicted with 90% reduction of bearing and eddy-current losses.

The above research studies have demonstrated the potential and feasibility of slotted AMM stators to be used to build high-efficiency AFPM machines. In terms of manufacturing, the cutting technique developed by the industrial partner has the potential for mass production of low-cost AMM machines. Furthermore, the research work presented here provides useful design guidelines. Hence, it is believed that this work will contribute to improving the design of slotted AMM AFPM machines.

### 9.1.2 Research Contribution Statement

The major research contributions developed in this work can be grouped into five main areas.

1. **Comparison of iron loss of SI, SMC and AMM materials**
   
   An extensive series of tests were conducted on a range of ring core samples to determine the B-H curves and specific iron losses at different frequencies. It was found that at 50Hz coated AMM has the lowest iron loss followed by grain-oriented SI, uncoated AMM, non-oriented SI and SMC.

   For estimating the specific iron loss of a slotted core, compensation method 2 based on the effective weight calculation from FE simulation provides the most accurate result. Compensation method 1 based on an average measured flux density assumption avoids the need for FE analysis and still provides reasonable accuracy.

   Based on the analytical intralaminar and interlaminar eddy-current losses models, it was found that the calculated ratio of interlaminar to intralaminar eddy-current loss is independent of frequency and flux density, and is proportional to the ratio of the interlamination to intralamination conductivities and the square of the ratio of the cross-sectional core width to the lamination thickness.

2. **3D finite element analysis - AFPM modelling**

   This research has made extensive use of 3D FEA modelling for the design of AFPM machines. The AFPM machines were modelled using quarter, half and
full 3D finite-element models. The "patch mesh function" was used to link the stator and rotor meshes across the rotating boundary. A mesh size selection and optimisation procedure was introduced to reduce the simulation time.

The calculated finite-element parameters include induced back-EMF, tooth and yoke flux density, cogging torque, iron loss and inductance. Time-stepping circuit-coupled simulations were performed to simulate operation with a six-step voltage source inverter. An extensive validation of the simulation results with the experimental results was performed and a correspondence of within 5 to 10% was generally obtained.

3. AMM AFPM machine design and analysis

A design procedure was proposed to give guidelines in determining the stator diameter, slot and pole combination, magnet thickness, slot depth, slot width, stator yoke thickness and winding turns of the machine taking into consideration of the limitations of the AMM cutting technique and its saturation value. The optimum inner/outer diameter ratio is determined based on the analytical torque production principles. The effect of changing the slot depth and slot width (affecting the magnetic and electric loading) were discussed.

4. Extensive testing of the AFPM AMM prototype machines

Extensive experimental measurements were taken on small (32 mm) and large (110 mm) diameter AMM prototype machines based on a fractional-slot concentrated -winding single-sided configuration. The non-optimised construction of the AMM stator due to non-insulated AMM strip and hence higher iron loss resulted in a lower efficiency. The AMM machine can offers higher efficiency with some optimisation.

It was demonstrated that copper loss is the main loss at low speed but at high speed the open-circuit loss becomes the dominating loss. It was observed that the 32 mm machine is more sensitive to the large axial force in the single-sided configuration which increases the bearing loss. For the 110 mm machine, it was found that the main contribution of open-circuit loss is the bearing loss due to the radial load and the high eddy-current losses in the sintered magnets and solid back-iron.

It was concluded that the high open-circuit loss was primarily due to the bearing and the rotor eddy-current losses.
5. Open-circuit loss analysis of single-sided axial-field PM machines

The four main components of the open-circuit loss of axial-field PM machines are the bearing, windage, stator iron and rotor eddy-current losses. The bearing and windage loss were separated experimentally and the stator iron and rotor eddy-current losses were determined with 3D FE simulations.

9.2 Future Work

This research work has demonstrated the potential and feasibility of slotted AMM stators in high efficiency AFPM machines. It has performed design, analysis and testing of slotted AMM axial-field permanent magnet machines. The research areas that can be addressed in the future studies are:

- Use of coated AMM materials in the stator to obtain iron loss compared to the uncoated version, which requires special attention.

- Effect of the winding and slot cutting process on the iron loss. This includes the increased iron loss due to the high tension during winding and the stress induced by the cutting process which could be reduced by a proper annealing process. As high tension is required to achieve high stacking factor, some tradeoffs are expected.

- Means for rotor eddy-current loss reduction in the rotor magnets and the back-iron. This includes using segmented magnets and laminated back-iron.

- Use of bonded magnets and a double-sided configuration to reduce the rotor eddy-current loss and the bearing loss due to axial forces.

- Refinement and extension of the analytical tool for more accurate modelling. This includes rotor eddy-current loss and thermal modelling.

- Development of means for optimising the machine designs by varying parameters including inner and outer diameters, slot depth and slot width.